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Commentary

THE Post Office Telephone service has received a great deal of publicity in the last year or so, and for two very different reasons. The first is the rapid strides which have recently taken place in the more complete mechanization of an already highly mechanized service; this is largely due to the introduction of electronic equipment, a notable example being the introduction of STD (Subscriber Trunk Dialling) at Bristol last December. The second reason is the Post Office's avowed intention to make this highly mechanized system more human: or in the words of the Postmaster General, 'It is to make sure that in a system where automatic machines will play an increasing part the customer can always get friendly personal service when he wants it'.

The first reason is obviously the prerogative of the engineer and, increasingly, the electronic engineer while the second, although non-technical is a matter which can, and should, be of interest to the engineer for, with any equipment which is intended for the usage of the public, the customer's convenience and the service which he obtains is of paramount importance. It is of little use for the engineer to evolve a complicated piece of equipment, or a system, if it fails to find ready acceptance by the intended user. It is largely to overcome these problems that the Post Office is to instigate this drive to make its services more personal.

The main inspiration for this 'humanizing' policy is the 'Ray' report 'Telephone Service and the Customer' which is the report of a team of Post Office officials and trade union representatives led by Mr. F. J. Ray, Director of Inland Telephone Communications, on a visit to America to study their telephone system. The team make it clear that in so short a visit (two weeks) they could not pretend to have gained more than a superficial impression of the system; nonetheless, there is a great deal of interest in the report and a great deal which could be beneficially applied on a far wider front than the telephone service.

The structure of the Bell Telephone System, is of course very different from that of the General Post Office. Heading the organization is the American Telephone and Telegraph Company which provides advice and assistance to the Bell Operating Companies of which there are twenty-two. In parallel with this is the Western Electric Company and the Bell Telephone Laboratories. The former carries out the manufacturing, purchasing, distributing and central office (telephone exchange) installation for the Bell System, while the latter carries out research and development work for the whole of the Bell System. While the A.T. & T. Co. does not exercise direct control over the individual companies, in the majority of cases it holds most, if not all, of their issued shares. It is, by any standards, a vast and up-to-date organization. At the end of 1957 the Bell System contained 52 000 000 telephone stations of which 92 per cent were connected to automatic exchanges. Of the total stations 28 per cent had some Direct Distance Dialling facilities (the American equivalent of STD), while 10 per cent could dial on a nationwide basis. The total staff engaged in operating the system is approximately 700 000 of whom 230 000 are telephone operating staff and supervisors.

The Ray report states that throughout this enormous structure there is one outstanding characteristic which underlies everything the System does; it is the desire to please the customer. And not only does the System plan to please the customer, it goes out of its way to help him to use the service he gets.

This desire to please starts right from the planning stage. When a particular line of development has been decided upon, at the earliest possible stage in the actual planning a substantial field trial is started, this is particularly so in the case of apparatus with which the customer will come into contact. Having obtained this direct evidence of customer reaction at an early stage in planning the System attaches great weight to it in coming to its final conclusions. When new types of apparatus or new facilities are finally introduced the Operating Companies go to great trouble to explain to the subscribers how they work and how they can obtain the utmost service from them.

This facet of the Bell System finds special scope in the repair service. Those who use this service are usually suffering from a sense of grievance and the Bell System aims at removing this feeling by the prompt and personal way in which fault report calls are handled as well as the speed with which the faults themselves are cleared.

By this desire and effort to please the customer the Bell Telephone System, according to the Ray report, has built up a great and enviable reputation and while much of this is far removed from engineering it points a lesson which can be learned with profit by all who design equipment or plan services for the use of others.

Linear Multi-Tapped Potentiometers with Loaded Outputs

By K. C. Garner*. B.Sc.(Eng.), A.M.I.E.E., A.F.R.Ae.S.

An analysis of multi-tapped linear potentiometers is given for the type where shunt resistors are connected between adjacent tapping points, and in the presence of an output load resistance.

Using the relationships of a generalized network with appropriate numerical substitutions obtained from the particular potentiometer arrangement being considered, evaluation is simple and systematic.

A procedure for designing potentiometers to generate non-linear functions is suggested.

(Voir page 247 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 253)

ONE of many ways of generating analogues of non-linear laws uses servo-driven linear potentiometers with fixed-valued shunt resistors connected across sections of their windings. The method is well known[1,2,3], but it is thought that a useful contribution to the technique would be to develop a systematic approach to the design of the networks; i.e. to the choice of shunt resistor values and tapping points. Considerable generalization is possible and although algebraically some of the relationships appear rather cumbersome, the method is quite simple and straightforward when used for design purposes, and numerical substitutions are inserted.

The analysis includes the practical consideration of the load across the potentiometer output. In fact the load resistance can often be turned to good effect to improve the accuracy of fit to the desired law.

The method used in this article is complementary to the diakoptic methods due to Kron and suggested by Shen[3], and for most purposes will prove equally direct for the restricted type of network considered.

The Four Basic Cases

Fig. 1 shows the typical circuit of a multi-tapped potentiometer of resistance R, having $m - 1$ tappings at distances $n_1, n_2, \ldots, n_m$, and with fixed valued resistors $R_1, R_2, \ldots, R_m$ connected between them. The wiper is at a distance K from the 'earthy' end, where K can vary between zero and unity. The input voltage is V_1 and the output voltage is V_o. The output load resistance is defined as R_L.

As K increases from zero to unity an equivalent network can be drawn each time the wiper crosses a tapping point and traverses a particular section of the potentiometer. These equivalent networks fall into three fundamental cases as illustrated by Fig. 2(a), (b) and (d). A fourth case, shown in Fig. 3(a), is applicable if a shunt resistor has been omitted and the wiper traverses this un-shunted section.

CASE 1 (*where* $0 \leqslant K \leqslant n_1$)

In this case, illustrated by Fig. 2(a), the portion of the potentiometer between the input and the tapping point at n_1 consists of the sum of all the paralleled resistances and sections of the potentiometer winding involved.

CASE 2 (*where* $n_1 + n_2 + \ldots + n_{x-1} \leqslant K \leqslant n_1 + n_2 \ldots + n_x$ *and where* $2 \leqslant x \leqslant m - 1$).

Fig. 2(b) shows the circuit for Case 2 where the wiper is traversing any section, except for the section connected directly to earth or the section connected directly to the input end.

This circuit can be drawn in an alternative bridge network form as in Fig. 2(c) where

R_a = Sum of the paralleled resistances for values of n_p where $m > p > x$

$R_b = [(n_1 + n_2 \ldots + n_x) - K]\,R$

$R_c = R_L$

R_d = Sum of the paralleled resistances for values of n_q where $0 < q \leqslant (x - 1)$

$R_e = R_x$

$R_f = [K - (n_1 + n_2 \ldots + n_{x-1})]\,R$

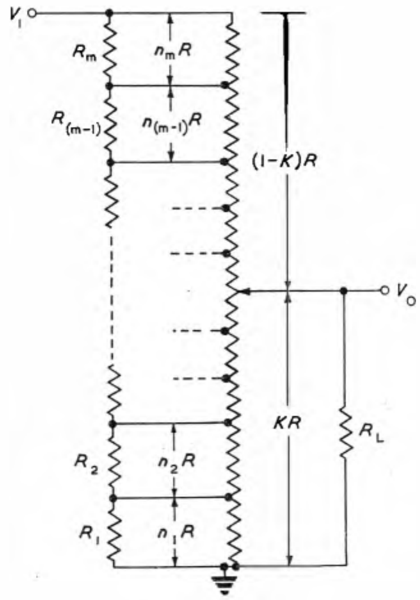

Fig. 1. Typical multi-tapped potentiometer circuit

CASE 3 (*where* $n_1 + n_2 + \ldots + n_{m-1} \leqslant K \leqslant 1$)

This circuit is given in Fig. 2(d) and as in Case 2 the alternative bridge network form of the circuit is shown in Fig. 2(e).

CASE 4 (*Wiper traversing an un-shunted section*)

If one or more fixed-valued shunt resistors are omitted between tappings a fourth case emerges when the wiper traverses such sections. Fig. 3(a) shows a typical example of this case, and Fig. 3(b) the equivalent circuit.

The Generalized Network

From further consideration of the circuits it appears that the network of Case 2 is in fact the generalized network, since it is possible to derive cases 1, 3 and 4 from it by

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making certain resistances zero or infinite. Thus case 1 may be obtained by considering $R_d = 0$. Case 3 is obtained by letting R_a vanish, and Case 4 by making $R_e = \infty$.

This approach to the problem enables one general equation, representing the attenuation of the bridge network of Fig. 2(c) to be used for all cases. By compiling the appropriate set of substitutions for each section (which are unique for the section being traversed by the wiper), the complete law of the multi-tapped potentiometer may be calculated.

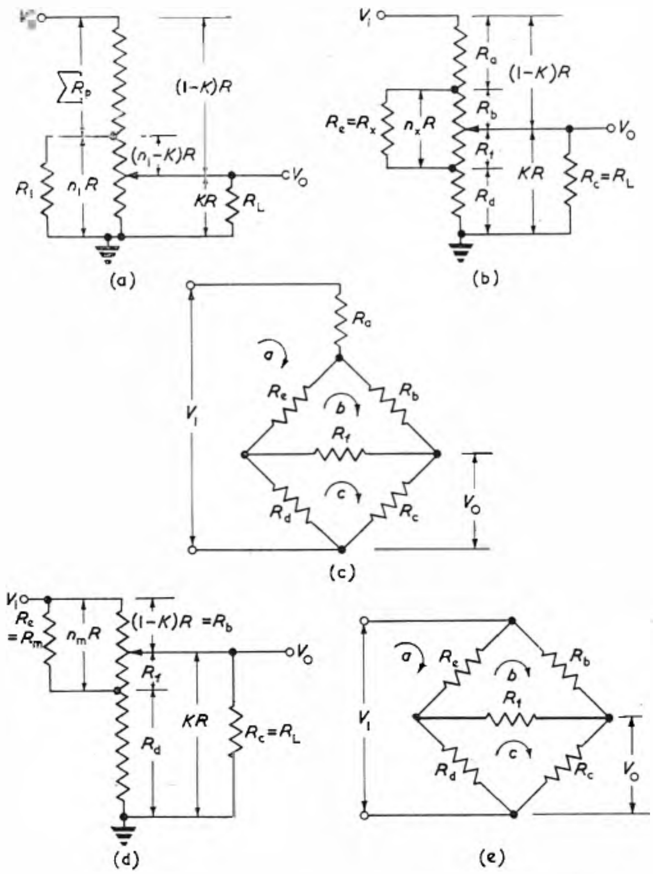

Fig. 2 (a). Equivalent circuit for Case 1

$$\sum R_p = \left\{ \begin{array}{c} \text{Sum of paralleled resistors from} \\ x=2 \text{ to } x=m \text{ where } m = \text{total number} \\ \text{of shunt resistors} \end{array} \right\} = \sum_{x=2}^{x=m} \frac{n_x R R_x}{(n_x R + R_x)}$$

(b) Equivalent circuit for Case 2

(c) Bridge form of the network of Fig. 2(b) (Case 2)

(d) Equivalent circuit for Case 3

R_d Sum of paralleled resistors for n_q where

$0 < q \leqslant m-1$

$R_f = [K-(n_1+n_2 \;.\;.\;.\; +n_{(m-1)})]\, R$

(e) Bridge form of Case 3

Analysis of the Generalized Network

The generalized network is shown in Fig. 2(c). Let the circulating currents a, b, and c be positive in the sense denoted by the arrows.

By inspection of the figure

$$b(R_b + R_f + R_e) = aR_e + cR_f = bZ_1 \quad \text{........ (1)}$$

$$c(R_c + R_d + R_f) = aR_d + bR_f = cZ_2 \quad \text{........ (2)}$$

and

$$V_1 = a(R_a + R_e + R_d) - cR_d - bR_e \quad \text{........ (3)}$$

or

$$V_1 = aZ_3 - cR_d - bR_e \quad \text{............ (4)}$$

from which

$$a = \frac{c(Z_1Z_2 - R_f^2)}{(Z_1R_d + R_eR_f)} \quad \text{............. (5)}$$

$$b = \frac{c(Z_2R_e + R_dR_f)}{(Z_1R_d + R_eR_f)} \quad \text{............. (6)}$$

Putting these values for a and b in equation (4) and rearranging it is found that

$$c = \frac{(Z_1R_d - R_eR_f)}{(Z_1Z_2Z_3 - Z_1R_d^2 - Z_2R_e^2 - Z_3R_f^2 - 2R_dR_eR_f)}\, V_1 \quad \text{.. (7)}$$

and remembering that the output load is represented by R_c one can write

$$V_o/V_1 = cR_c/V_1 \quad \text{............. (8)}$$

so that the general equation is*

$$V_o/V_1 = \frac{R_c(Z_1R_d + R_eR_f)}{(Z_1Z_2Z_3 - Z_1R_d^2 - Z_2R_e^2 - Z_3R_f^2 - 2R_dR_eR_f)}. \quad (9)$$

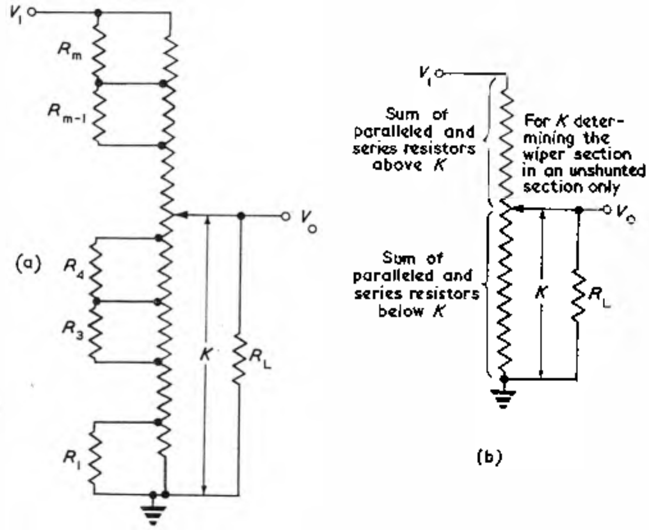

Fig. 3(a). Potentiometer network with resistors omitted between tappings (Case 4) (when wiper traverses an unshunted section)

(b) Equivalent circuit for Case 4

where $Z_1 = (R_b + R_f + R_e)$ .... (10)

$Z_2 = (R_c + R_d + R_f)$ .... (11)

$Z_3 = (R_a + R_e + R_d)$ .... (12)

Analysis of Each of the Four Cases

Case 1 $(0 \leqslant K \leqslant n_1)$

As stated above, Case 1 is obtained by making R_d vanish in equation (9) with the result that the equation then represents the network of Fig. 2(a) with the appropriate values substituted. Thus the characteristic is given by

$$V_o/V_1 = \frac{R_c(Z_1R_d + R_eR_f)}{(Z_1Z_2Z_3 - Z_1R_d^2 - Z_2R_e^2 - Z_3R_f^2 - 2R_dR_eR_f)}. \quad (13)$$

where

$$Z_1 = (R_b + R_f + R_e) = (n_1 - K)R + KR + R_1 = (n_1R + R_1) \quad \text{............ (14)}$$

$$Z_2 = (R_c + R_d + R_f) = (R_L + KR) \quad \text{........ (15)}$$

$$Z_3 = (R_a + R_e + R_d) = (\textstyle\sum R_p + R_1) \quad \text{...... (16)}$$

$R_a = \sum R_p$ = Sum of paralleled resistances above tapping at n_1 (i.e. for values of n_p where $m \geqslant p > 1$) .... (17)

* The expression in this form exhibits certain redundancies apparent from the minus signs and squared terms. It is, however, considered to be in the easiest form to memorize.

$$R_b = (n_1 - K)R \quad \ldots\ldots \quad (18)$$
$$R_c = R_L \quad \ldots\ldots \quad (19)$$
$$R_d = 0 \quad \ldots\ldots \quad (20)$$
$$R_e = R_1 \quad \ldots\ldots \quad (21)$$
$$R_f = KR \quad \ldots\ldots \quad (22)$$

CASE 2 ($n_1 + n_2 + \ldots + n_{x-1} \leqslant K \leqslant n_1 + n_2 + \ldots + n_x$) and where ($2 \leqslant x \leqslant m - 1$)

This case has been illustrated in Fig. 2(b), but for the purposes of analysis it is preferable to consider Fig. 2(c). The relationship already obtained is

$$V_o/V_1 = \frac{R_c(Z_1R_d + R_eR_f)}{(Z_1Z_2Z_3 - Z_1R_d^2 - Z_2R_e^2 - Z_3R_f^2 - 2R_dR_eR_f)} \quad (23)$$

with the following substitutions

R_a = Sum of the paralleled resistances for values of n_p where $m \geqslant p > x$ ...... (24)

$$R_b = [(n_1 + n_2 + \ldots + n_x) - K]R \quad \ldots\ldots \quad (25)$$
$$R_c = R_L \quad \ldots\ldots \quad (26)$$

R_d = Sum of the paralleled resistances for values of n_q where $0 < q \leqslant (x - 1)$ ...... (27)

$$R_e = R_x \quad \ldots\ldots \quad (28)$$
$$R_f = [K - (n_1 + n_2 + \ldots + n_{x-1})]R \quad \ldots\ldots \quad (29)$$
and
$$Z_1 = (R_b + R_f + R_e) \quad \ldots\ldots \quad (30)$$
$$Z_2 = (R_c + R_d + R_f) \quad \ldots\ldots \quad (31)$$
$$Z_3 = (R_a + R_e + R_d) \quad \ldots\ldots \quad (32)$$

CASE 3 ($n_1 + n_2 + \ldots + n_{m-1} \leqslant K \leqslant 1$)

Case 3 is given in Fig. 2(d) and the bridge form shown in Fig. 2(e). Again the network attenuation is obtained using the generalized equation

$$V_o/V_1 = \frac{R_c(Z_1R_d + R_eR_f)}{(Z_1Z_2Z_3 - Z_1R_d^2 - Z_2R_e^2 - Z_3R_f^2 - 2R_dR_eR_f)} \quad (33)$$

but with the following substitutions in this case:

$$R_a = 0 \quad \ldots\ldots \quad (34)$$
$$R_b = (1 - K)R \quad \ldots\ldots \quad (35)$$
$$R_c = R_L \quad \ldots\ldots \quad (36)$$

R_d = Sum of the paralleled resistances for values of n_q where $0 < q \leqslant (m - 1)$ ...... (37)

$$R_e = R_m \quad \ldots\ldots \quad (38)$$
$$R_f = [K - (n_1 + n_2 + \ldots + n_{m-1})]R \quad \ldots\ldots \quad (39)$$
and
$$Z_1 = (R_b + R_f + R_e) \quad \ldots\ldots \quad (40)$$
$$Z_2 = (R_c + R_d + R_f) \quad \ldots\ldots \quad (41)$$
$$Z_3 = (R_a + R_e + R_d) \quad \ldots\ldots \quad (42)$$

CASE 4 (*when shunt resistors are omitted between tappings, and the wiper is traversing an un-shunted section*)

Fig. 3(b) illustrates this case, and it is evident that the same general equation holds good provided that $R_e = \infty$. Thus

$$V_o/V_1 = \frac{R_c(Z_1R_d + R_eR_f)}{(Z_1Z_2Z_3 - Z_1R_d^2 - Z_2R_e^2 - Z_3R_f^2 - 2R_dR_eR_f)} \quad (43)$$

where

$$Z_1 = (R_b + R_e + R_f) \quad \ldots\ldots \quad (44)$$
$$Z_2 = (R_c + R_d + R_f) \quad \ldots\ldots \quad (45)$$
$$Z_3 = (R_a + R_d + R_e) \quad \ldots\ldots \quad (46)$$

and $R_e = \infty$

so that neglecting all terms except those containing R_e

$$V_o/V_1 = \frac{(R_d + R_f)R_cR_e}{(R_a + R_b)(R_c + R_d + R_f)R_e + (R_d + R_f)R_cR_e} \quad (47)$$

or

$$V_o/V_1 = \frac{(R_d + R_f)R_c}{(R_a + R_b)(R_c + R_d + R_f) + (R_d + R_f)R_c} \quad \ldots \quad (48)$$

where

R_a = Sum of the paralleled and series resistances for values of n_p where $m \geqslant p > x$ ...... (49)

$$R_b = [(n_1 + n_2 + \ldots + n_x) - K]R \quad \ldots\ldots \quad (50)$$
$$R_c = R_L \quad \ldots\ldots \quad (51)$$

R_d = Sum of the paralleled and series resistances for values of n_q where $0 < q \leqslant (x - 1)$ ...... (52)

$$R_f = [K - (n_1 + n_2 + \ldots + n_{x-1})]R \quad \ldots\ldots \quad (53)$$

The Application of the Analysis to Typical Networks

From the relationships derived so far, all the arrangements of multi-tapped potentiometers of the type shown in Fig. 1 can be evaluated by choosing suitable values for

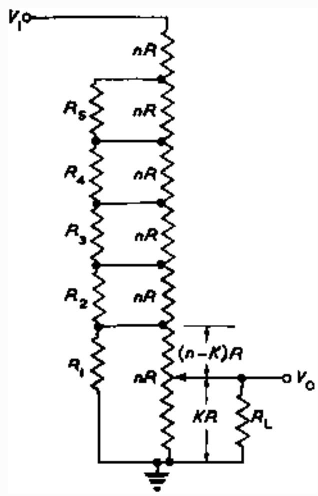

Fig. 4. A multi-tapped potentiometer with an output load (Case 1)

R_a, R_b, R_c, R_d, R_e and R_f. In the more simple arrangements considered later it is probably more convenient, analytically, to derive the characteristic from first principles. However, when direct numerical substitutions are made in the generalized equation the calculation is quite simple.

A MULTI-TAPPED POTENTIOMETER WITH AN OUTPUT LOAD

To illustrate this example a potentiometer has been chosen having five tapping points spaced at equal distances, n, along the winding. These tapping points are numbered in such a way that the first is at a distance n from the 'earthy' terminal, and the fifth is at a distance n from the input terminal. These tapping points are connected by resistors R_1, R_2, R_3, R_4 and R_5. There is no resistor between the fifth tapping point and the input terminal.

Fig. 4 shows the potentiometer with an output load R_L, and with the wiper in the first section under the condition $0 \leqslant K \leqslant n$, so that the resulting network is an example of Case 1 and the substitution values are:

$$R_a = \frac{nRR_2}{(nR + R_2)} + \frac{nRR_3}{(nR + R_3)} + \frac{nRR_4}{(nR + R_4)} + \frac{nRR_5}{(nR + R_5)} + nR \quad \ldots \quad (54)$$

$$= nR\left[\frac{1}{(n\rho_2 + 1)} + \frac{1}{(n\rho_3 + 1)} + \frac{1}{(n\rho_4 + 1)} + \frac{1}{(n\rho_5 + 1)} + 1\right] \quad \ldots\ldots \quad (55)$$

and letting such terms as $\frac{1}{n\rho_x + 1}$ be represented by γ_x then

$$R_a = nR(\gamma_2 + \gamma_3 + \gamma_4 + \gamma_5 + 1) \quad \ldots\ldots \quad (56)$$

$$R_b = (n - K)R \quad (57)$$
$$R_c = R_L \quad (58)$$
$$R_d = 0 \quad (59)$$
$$R_e = R_1 \quad (60)$$
$$R_f = KR \quad (61)$$

so that for $(0 \leqslant K \leqslant n)$..

$$V_o/V_1 = \frac{K\gamma_1}{n(1+\gamma_1+\gamma_2+\gamma_3+\gamma_4+\gamma_5)(1+K\rho_L-K^2\rho_L\rho_1\gamma_1)-K^2\rho_L\gamma_1^2} \quad (62)$$

The case when $\rho_L = 0$ is given by

$$V_o/V_1 \text{ (no load)} = \frac{K\gamma_1}{n(\gamma_1+\gamma_2+\gamma_3+\gamma_4+\gamma_5+1)} \quad (63)$$

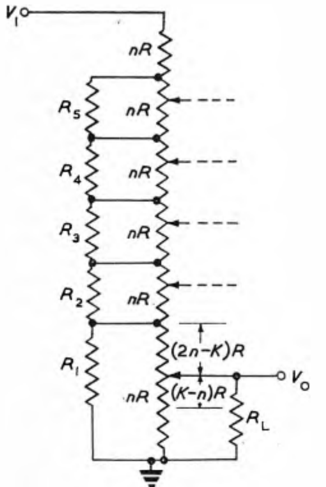

Fig. 5. A multi-tapped potentiometer with an output load (Case 2)

The voltage when the wiper is on the first tapping point is given by

$$V_o/V_1 \text{ (for } K = n) = \frac{1}{(\gamma_2+\gamma_3+\gamma_4+\gamma_5+1)(n\rho_1+n\rho_L+1)+1} \quad (64)$$

and

$$V_o/V_1 \text{ (no load and } K = n) = \frac{\gamma_1}{(\gamma_1+\gamma_2+\gamma_3+\gamma_4+\gamma_5+1)} \quad (65)$$

When the wiper traverses $(n \leqslant K \leqslant 2n)$, $(2n \leqslant K \leqslant 3n)$, $(3n \leqslant K \leqslant 4n)$, $(4n \leqslant K \leqslant 5n)$, the networks formed in each condition are examples of Case 2. A set of substitutions are obtained from Fig. 5 to form the equation for each mode of operation and are listed in Table 1.

TABLE 1

RESISTANCE NOTATION	$n \leqslant K \leqslant 2n$	$2n \leqslant K \leqslant 3n$	$3n \leqslant K \leqslant 4n$	$4n \leqslant K \leqslant 5n$
R_a	$nR(1+\gamma_5+\gamma_4+\gamma_3)$	$nR(1+\gamma_5+\gamma_4)$	$nR(1+\gamma_5)$	nR
R_b	$(2n-K)R$	$(3n-K)R$	$(4n-K)R$	$(5n-K)R$
R_c	R_L	R_L	R_L	R_L
R_d	$nR\gamma_1$	$nR(\gamma_1+\gamma_2)$	$nR(\gamma_1+\gamma_2+\gamma_3)$	$nR(\gamma_1+\gamma_2+\gamma_3+\gamma_4)$
R_e	R_2	R_3	R_4	R_5
R_f	$(K-n)R$	$(K-2n)R$	$(K-3n)R$	$(K-4n)R$

If these substitutions are given their numerical values the attenuation characteristic can be calculated for the condition when there is an output load. When designing a non-linear function generator, however, the most useful expressions, in the first instance, are those giving the no-load condition. These can be obtained by using the above substitutions and making $R_c = R_L = 0$, so that

For $(n \leqslant K \leqslant 2n)$

$$V_o/V_1 \text{ (no load)} = \frac{(K-n)\gamma_2+n\gamma_1}{n(\gamma_1+\gamma_2+\gamma_3+\gamma_4+\gamma_5+1)} \quad (66)$$

For $(2n \leqslant K \leqslant 3n)$

$$V_o/V_1 \text{ (no load)} = \frac{(K-2n)\gamma_3+n(\gamma_1+\gamma_2)}{n(\gamma_1+\gamma_2+\gamma_3+\gamma_4+\gamma_5+1)} \quad (67)$$

For $(3n \leqslant K \leqslant 4n)$

$$V_o/V_1 \text{ (no load)} = \frac{(K-3n)\gamma_4+n(\gamma_1+\gamma_2+\gamma_3)}{n(\gamma_1+\gamma_2+\gamma_3+\gamma_4+\gamma_5+1)} \quad (68)$$

For $(4n \leqslant K \leqslant 5n)$

$$V_o/V_1 \text{ (no load)} = \frac{(K-4n)\gamma_5+n(\gamma_1+\gamma_2+\gamma_3+\gamma_4)}{n(\gamma_1+\gamma_2+\gamma_3+\gamma_4+\gamma_5+1)} \quad (69)$$

and hence the voltage on the wiper when in contact with the various tappings points in the absence of the load resistance is

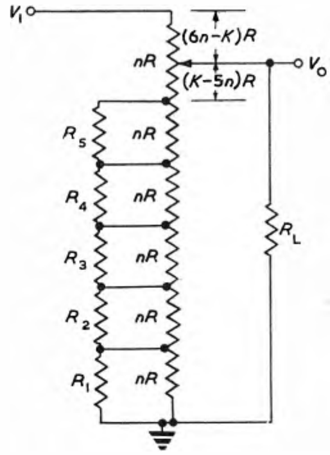

Fig. 6. A multi-tapped potentiometer with an output load (Case 4)

$$V_o/V_1 \text{(no load, } K = n) = \frac{\gamma_1}{(\gamma_1+\gamma_2+\gamma_3+\gamma_4+\gamma_5+1)} \quad (70)$$

$$V_o/V_1 \text{(no load, } K = 2n) = \frac{(\gamma_1+\gamma_2)}{(\gamma_1+\gamma_2+\gamma_3+\gamma_4+\gamma_5+1)} \quad (71)$$

$$V_o/V_1 \text{ (no load, } K = 3n) = \frac{(\gamma_1+\gamma_2+\gamma_3)}{(\gamma_1+\gamma_2+\gamma_3+\gamma_4+\gamma_5+1)} \quad (72)$$

$$V_o/V_1 \text{ (no load, } K = 4n) = \frac{(\gamma_1+\gamma_2+\gamma_3+\gamma_4)}{(\gamma_1+\gamma_2+\gamma_3+\gamma_4+\gamma_5+1)} \quad (73)$$

$$V_o/V_1 \text{ (no load, } K = 5n) = \frac{(\gamma_1+\gamma_2+\gamma_3+\gamma_4+\gamma_5)}{(\gamma_1+\gamma_2+\gamma_3+\gamma_4+\gamma_5+1)} \quad (74)$$

When the wiper moves in the section $5n \leqslant K \leqslant 1$ the network becomes an example of Case 4 and the following substitutions obtained from Fig. 6 are used:—

$$R_a = 0 \quad (75)$$
$$R_b = (6n - K)R \quad (76)$$
$$R_c = R_L \quad (77)$$
$$R_d = nR(\gamma_1+\gamma_2+\gamma_3+\gamma_4+\gamma_5) \quad (78)$$
$$R_e = \infty \quad (79)$$
$$R_f = (K - 5n)R \quad (80)$$

With these values the following relationship is obtained:

$$V_o/V_1 = \frac{(K-5n)+n(\gamma_1+\gamma_2+\gamma_3+\gamma_4+\gamma_5)}{(6n-K)[(K-5n)\rho_L+n\rho_L(\gamma_1+\gamma_2+\gamma_3+\gamma_4+\gamma_5)+1]+(K-5n)+n(\gamma_1+\gamma_2+\gamma_3+\gamma_4+\gamma_5)} \quad (81)$$

and the value for no load, (i.e. $\rho_L = 0$) is

$$V_o/V_1\,(\text{no load}) = \frac{(K-5n)+n(\gamma_1+\gamma_2+\gamma_3+\gamma_4+\gamma_5)}{n(\gamma_1+\gamma_2+\gamma_3+\gamma_4+\gamma_5+1)} \quad (82)$$

The Simple Linear Potentiometer with an Output Load

This circuit is included for the sake of completeness, and for the importance of the result in determining the loading error of linear servo multipliers. The circuit is shown in Fig. 7. Since it represents the case of a completely un-shunted 'multi-tapped' potentiometer it can be considered as an example of Case 4.

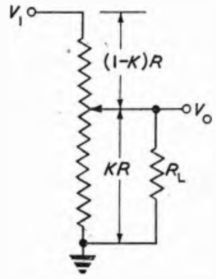

Fig. 7. Linear potentiometer with output load

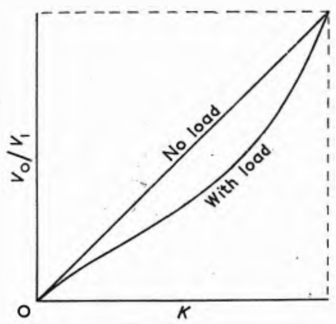

Fig. 8. Characteristic of loaded linear potentiometer

The substitutions for this network are

$$R_a = 0 \quad (83)$$
$$R_b = (1-K)R \quad (84)$$
$$R_c = R_L \quad (85)$$
$$R_d = 0 \quad (86)$$
$$R_e = \infty \quad (87)$$
$$R_f = KR \quad (88)$$

so that

$$V_o/V_1 = \frac{KRR_L}{(1-K)R(R_L+KR)+KRR_L} \quad (89)$$

or

$$V_o/V_1 = \frac{K}{1+K\rho_L(1-K)} \quad (90)$$

i.e. the law of a practical servo-driven multiplying potentiometer, where $\rho_L = (R/R_L)$

In the absence of the load resistance $R_L, \rho_L = 0$ so that

$$V_o/V_1\,(\text{no load}) = K \quad (91)$$

i.e. an ideal servo-driven multiplying potentiometer.

Fig. 8 shows a typical characteristic curve for this network.

The Single-Tapped Potentiometer with an Output Load

Some non-linear functions can be generated to an acceptable accuracy using a standard centre-tapped loaded linear potentiometer and the following results are applicable for networks derived this way.

There are two variations of this circuit:

(a) Type 1, when the shunt resistor is taken to the "earthy" end.

(b) Type 2, when the shunt resistor is taken to the input end.

A third type is possible where there are two shunt resistors, but this can be resolved into networks of the forms of Cases 2 and 3.

Type 1.

The circuit for Type 1 is shown in Fig. 9 (a) and (b). Fig. 9(a) represents the condition for Case 1 ($0 \leqslant K \leqslant n$), and Fig. 9(b) that for Case 4 ($n \leqslant K \leqslant 1$). The circuit is such that the shunt resistor R_1 is connected between the tapping point and earth, and a load resistance R_L is connected between the wiper and earth.

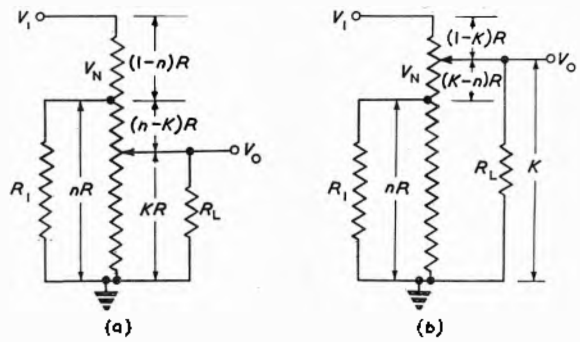

Fig. 9. Single-tapped potentiometer with fixed resistor connected from tapping to earth with an output load (Type 1)
(a) Case 1; (b) Case 4

For Case 1 ($0 \leqslant K \leqslant n$) the following substitutions are obtained by inspection of the Fig. 9(a):

$$R_a = (1-n)R \quad (92)$$
$$R_b = (n-K)R \quad (93)$$
$$R_c = R_L \quad (94)$$
$$R_d = 0 \quad (95)$$
$$R_e = R_1 \quad (96)$$
$$R_f = KR \quad (97)$$

and by substituting these values into the generalized equation

$$V_o/V_1 = \frac{KRR_LR_1}{(nR+R_1)(KR+R_L)[(1-n)R+R_1]-[(1-n)R+R_1]K^2R^2-(KR+R_L)R_1^2} \quad (98)$$

$$= \frac{K}{Kn\rho_1\rho_L - Kn^2\rho_1\rho_L + n\rho_1 - n^2\rho_1 + K\rho_L + 1 - K^2\rho_1\rho_L + K^2n\rho_1\rho_L - K^2\rho_L} \quad (99)$$

or alternatively by rearranging

$$V_o/V_1 = \frac{K}{1+K\rho_L(1-K)+\rho_1(1-n)[n+K\rho_L(n-K)]} \quad (100)$$

where $\rho_1 = (R/R_1)$ and $\rho_L = (R/R_L)$

For the no-load condition $\rho_L = 0$ so that

$$V_o/V_1\,(\text{no load}) = \frac{K}{1+n\rho_1(1-n)} \quad (101)$$

For Case 4 ($n \leqslant K \leqslant 1$) a similar set of substitutions are obtained from Fig. 9(b) so that

$$R_a = 0 \quad (102)$$
$$R_b = (1-K)R \quad (103)$$
$$R_c = R_L \quad (104)$$
$$R_d = \frac{NRR_1}{(nR+R_1)} \quad (105)$$
$$R_e = \infty \quad (106)$$
$$R_f = (K-n)R \quad (107)$$

and since $R_e = \infty$, these values may be substituted in equation (48)

Thus

$$V_o/V_1 = \frac{\left[\frac{nRR_1}{(nR+R_1)} + (K-n)R\right]R_L}{(1-K)R\left[\frac{nRR_1}{(nR+R_1)} + (K-n)R + R_L\right] + \left[\frac{nRR_1}{(nR+R_1)} + (K-n)R\right]R_L} \quad \ldots (108)$$

or

$$V_o/V_1 = \frac{K}{1 + K\rho_L(1-K) + \frac{n^2\rho_1(1-K)}{n\rho_1(K-n)+K}} \quad \ldots\ldots (109)$$

where $\rho_1 = (R/R_1)$ and $\rho_L = (R/R_L)$

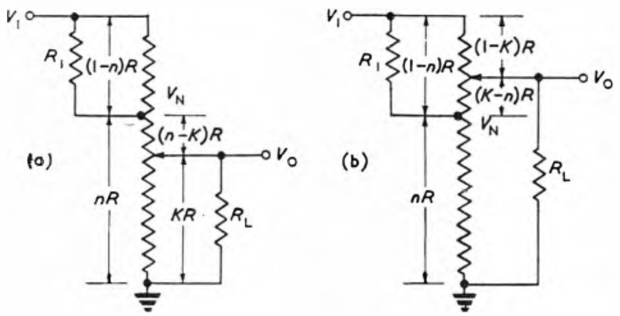

Fig. 10. Single-tapped potentiometer with fixed resistor connected from tapping to input with an output load (Type 2)
(a) Case 4 ; (b) Case 2

For the no-load condition $\rho_L = 0$, so that

$$V_o/V_1 \text{ (no load)} = \frac{K + n\rho_1(K-n)}{1 + n\rho_1(1-n)} \quad \ldots\ldots\ldots (110)$$

Type 2

Fig. 10(a) shows the circuit condition when $0 \leqslant K \leqslant n$, and is again an example of Case 4 since the wiper is traversing an un-shunted section. Therefore the following substitutions into equation (48) are made:

$$R_a = \frac{(1-n)RR_1}{(1-n)R+R_1} \quad \ldots\ldots\ldots\ldots (111)$$

$$R_b = (n-K)R \quad \ldots\ldots\ldots\ldots (112)$$

$$R_c = R_L \quad \ldots\ldots\ldots\ldots (113)$$

$$R_d = 0 \quad \ldots\ldots\ldots\ldots (114)$$

$$R_e = \infty \quad \ldots\ldots\ldots\ldots (115)$$

$$R_f = KR \quad \ldots\ldots\ldots\ldots (116)$$

so that

$$V_o/V_1 = \frac{KRR_L}{\left[\frac{(1-n)RR_1}{(1-n)R+R_1} + (n-K)R\right](KR+R_L) + KRR_L} \quad \ldots\ldots\ldots\ldots (117)$$

or

$$V_o/V_1 = \frac{K}{\left[\frac{(1-n)}{(1-n)\rho_1+1} + (n-K)\right](K\rho_L+1)+K} \quad \ldots (118)$$

where $\rho_1 = (R/R_1)$ and $\rho_L = (R/R_L)$

For the no load condition when $\rho_L = 0$

$$V_o/V_1 \text{ (no load)} = \frac{K[1+\rho_1(1-n)]}{1+n\rho_1(1-n)} \quad \ldots\ldots\ldots (119)$$

For the attenuation when $n \leqslant K \leqslant 1$, illustrated by Fig. 10(b) the substitutions are as follows:

$$R_a = 0 \quad \ldots\ldots\ldots\ldots (120)$$

$$R_b = (1-K)R \quad \ldots\ldots\ldots\ldots (121)$$

$$R_c = R_L \quad \ldots\ldots (122)$$

$$R_d = nR \quad \ldots\ldots (123)$$

$$R_e = R_1 \quad \ldots\ldots (124)$$

$$R_f = (K-n)R \quad \ldots (125)$$

which gives

$$V_o/V_1 = \frac{[nR(1-n)+KR_1]R_L}{[nR(1-n)+KR_1]R_L + [K(1-K+n)-n]nR^2 + KRR_1(1-K)} \quad \ldots\ldots\ldots\ldots (126)$$

or

$$V_o/V_1 = \frac{[n\rho_1(1-n)+K]}{[n\rho_1(1-n)+1] + n\rho_1\rho_L[K(1-K+n)-n] + K\rho_L(1-K)} \quad \ldots\ldots\ldots\ldots (127)$$

and the no-load condition when $\rho_L = 0$ is

$$V_o/V_1 \text{ (no load)} = \frac{K + n\rho_1(1-n)}{1+n\rho_1(1-n)} \quad \ldots\ldots (128)$$

Suggested Design Procedure when applied to the Generation of Non-Linear Functions

Using the analytical methods contained in this note, it is possible to design tapped-potentiometer networks to approximate many monotonic functions (e.g. secant law, logarithmic law, parabolic law, etc.). Basically, the problem resolves itself into determining,

(1) The number of tapping points which are necessary to ensure that the required accuracy is obtained.

(2) The values required for the shunt resistors.

(3) The best value for the output load, if this is controllable.

A possible procedure is outlined below.

Design Procedure

Given the non-linear law, either in graphical form or as a mathematical expression, the first step would be to estimate the number of tapping-points necessary to obtain the desired accuracy. This may be done by postulating a trial number of tapping points at known spacings and assuming that the voltages at the tapping-points can be made to lie on the non-linear function curve, in the absence of an output load. With the assumption that $R_L = \infty$, the slope of V_o/V_1 between tappings will be constant, so that by joining the tapping-points as they occur on the non-linear law curve, with straight lines, as in Fig. 11, the no-load curve of the potentiometer network may be drawn. From this curve the maximum error is readily determined. If this maximum error exceeds the permissible tolerance the number of tapping-points should be increased and/or the spacings decreased over parts of the non-linear curve where the radius of curvature is small. In practice the spacing can usually be made equal, provided enough tapping-points can be made available, with consequent simplification of the design calculations which are to follow.

Because generalization is difficult it will be assumed that, on the basis of the above criterion, a satisfactory result will be obtained by using the number of tapping-points at the spacings shown in the potentiometer network of Fig. 4, i.e. five tapping points each at equal distances apart. The next step is to determine the shunt resistor values which will satisfy the condition that the voltages at the tapping points, in the absence of an output load, will lie on the non-linear function curve. Such values will quite possibly be satisfactory for the final design in the presence of an output load, and will be of the right order.

Let the non-linear law be

$$\alpha = Cf_1(\beta) \quad \ldots\ldots\ldots\ldots\ldots\ (129)$$

or expressed as voltage analogues

$$V_o = V_1 f_1(K) \quad \ldots\ldots\ldots\ldots\ (130)$$

Since it has been assumed that the potentiometer-network of Fig. 4 is to be the basic network for the design, the

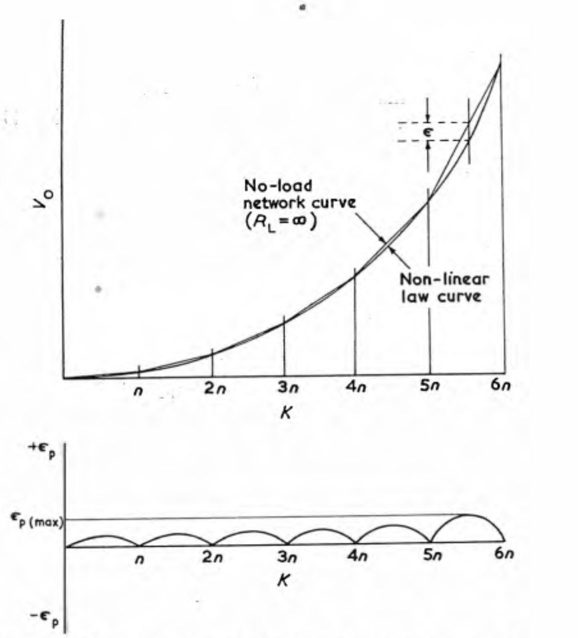

Fig. 11. No-load network curve and proportional error curve

values of the shunt resistors can be obtained by solving for R_1, R_2, R_3, R_4, and R_5 from equations (70) to (74) where V_o/V_1 (no load, $K=n$ to $5n$)$=(V_{X1\text{-}5}/V_1)=a_1$ to a_5 .. (131)

a_1 to a_5 being ratios derived from the law

$$V_o = V_1 f_1(K) \quad \ldots\ldots\ldots\ldots\ldots\ (132)$$

for values of $K = n$ to $5n$.

Thus

$$V_{X1}/V_1 = (\gamma_1/Z) = a_1 \quad \ldots\ldots\ldots\ (133)$$

$$V_{X2}/V_1 = \frac{(\gamma_1+\gamma_2)}{Z} = a_2 \quad \ldots\ldots\ldots\ (134)$$

$$V_{X3}/V_1 = \frac{(\gamma_1+\gamma_2+\gamma_3)}{Z} = a_3 \quad \ldots\ (135)$$

$$V_{X4}/V_1 = \frac{(\gamma_1+\gamma_2+\gamma_3+\gamma_4)}{Z} = a_4 \quad \ldots\ (136)$$

$$V_{X5}/V_1 = \frac{(\gamma_1+\gamma_2+\gamma_3+\gamma_4+\gamma_5)}{Z} = a_5 \quad \ldots\ (137)$$

where

$$Z = (\gamma_1+\gamma_2+\gamma_3+\gamma_4+\gamma_5+1) \quad \ldots\ldots\ (138)$$

Therefore

$$\gamma_1/Z = a_1 \quad \text{or } \gamma_1 = a_1 Z \quad \ldots\ldots\ldots\ldots\ (139)$$

$$\gamma_2/Z = (a_2 - a_1) \text{ or } \gamma_2 = (a_2 - a_1)Z \quad \ldots\ldots\ (140)$$

$$\gamma_3/Z = (a_3 - a_2) \text{ or } \gamma_3 = (a_3 - a_2)Z \quad \ldots\ldots\ (141)$$

$$\gamma_4/Z = (a_4 - a_3) \text{ or } \gamma_4 = (a_4 - a_3)Z \quad \ldots\ldots\ (142)$$

$$\gamma_5/Z = (a_5 - a_4) \text{ or } \gamma_5 = (a_5 - a_4)Z \quad \ldots\ldots\ (143)$$

Substituting these values in equation (137)

$$Z = \frac{1}{(1 - a_5)} \quad \ldots\ldots\ldots\ldots\ (144)$$

$$\gamma_1 = \frac{a_1}{(1 - a_5)} \quad \ldots\ldots\ldots\ldots\ (145)$$

$$\gamma_2 = \frac{(a_2 - a_1)}{(1 - a_5)} \quad \ldots\ldots\ldots\ldots\ (146)$$

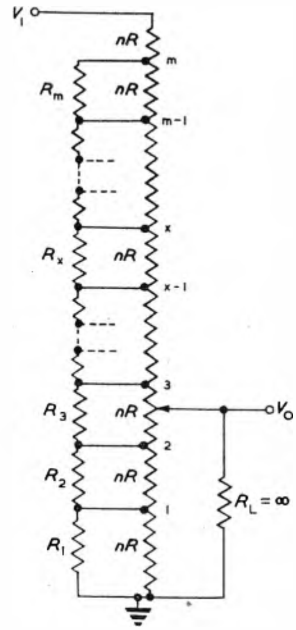

Fig. 12. Multi-tapped potentiometer with m equidistant taps

$$R_x = \frac{n(a_x - a_{x-1})R}{(1 - a_m - a_x + a_{x-1})}$$

where $a_0 = 0$

$$\gamma_3 = \frac{(a_3 - a_2)}{(1 - a_5)} \quad \ldots\ldots\ldots\ldots\ (147)$$

$$\gamma_4 = \frac{(a_4 - a_3)}{(1 - a_5)} \quad \ldots\ldots\ldots\ldots\ (148)$$

$$\gamma_5 = \frac{(a_5 - a_4)}{(1 - a_5)} \quad \ldots\ldots\ldots\ldots\ (149)$$

and by definition $\gamma_x = \dfrac{1}{n\rho_x + 1}$ and $\rho_x = (R/R_x)$ so that

$$R_1 = \frac{na_1 R}{(1 - a_5 - a_1)} \quad \ldots\ldots\ldots\ldots\ (150)$$

$$R_2 = \frac{n(a_2 - a_1)R}{(1 - a_5 - a_2 + a_1)} \quad \ldots\ldots\ldots\ (151)$$

$$R_3 = \frac{n(a_3 - a_2)R}{(1 - a_5 - a_3 + a_2)} \quad \ldots\ldots\ldots\ (152)$$

$$R_4 = \frac{n(a_4 - a_3)R}{(1 - a_5 - a_4 + a_3)} \quad \ldots\ldots\ldots\ (153)$$

$$R_5 = \frac{n(a_5 - a_4)R}{(1 - 2a_5 + a_4)} \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (154)$$

which by induction leads to the general result, that for equal spacings with m tapping points and no output load, and with no shunt resistor from the input terminal to the nearest tapping point (i.e. for potentiometers of the form shown in Fig. 12):

$$R_x = \frac{n(a_x - a_{x-1})R}{(1 - a_m - a_x + a_{x-1})} \quad \text{where } a_0 = 0 \ldots . \quad (155)$$

It can be seen that the value of R_x depends on R, the potentiometer winding resistance. This is because there is no shunt resistor between the m^{th} tapping point and the input terminal in the considered design. If there were, then the shunt resistor values would be independent of R, but in practice the resistor between the m^{th} tapping point and the input terminal should be omitted since it is essentially redundant and it reduces the input impedance of the network.

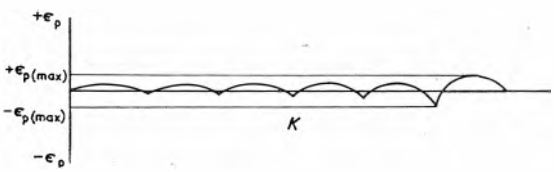

Fig. 13. Proportional error curve for loaded network

With the values for the shunt resistors determined by equations (150) to (154) the network will have (in the absence of an output load) the same maximum error as originally computed.

The next step is to compute the attenuation curve of the network in the presence of the minimum practical output load, e.g. the input impedance of the succeeding d.c. amplifier, using the generalized equation and the substitutions derived for Fig. 4. From these values a new curve can be plotted and compared with the non-linear law curve to determine the new maximum error.

Provided the minimum practical output load is small some improvement in accuracy should be found, since under no-load conditions only positive errors are present, and the effect of introducing an output load is to cause the straight-line portions of the no-load curve to bow slightly downwards at the centre, and the tapping point voltages to decrease slightly below the non-linear law curve, thus introducing compensating negative errors (see Fig. 13).

It can be seen that if the output load is reduced too far, the negative errors produced will become dominant. Thus for a given set of shunt resistors there is an optimum value for R_L. If the minimum practical output load is inherently greater than this optimum, more tapping-points must be used to improve the accuracy.

At this stage, however, the design may well be satisfactory, but if further improvement is required the output load R_L can be optimized. The optimum value for R_L is that which produces equal positive and negative percentage errors.

The process of optimizing R_L if the non-linear function is only known graphically must be a process of trial and error by drawing out curves of error versus R_L, and by studying their shape choosing an appropriate value.

If the function is known in analytical form then an analytical approach can be adopted although the method may be very tedious compared with a graphical method.

Whatever method is chosen to compute the resistors required in the final refined design, it must be borne in mind throughout the design stage that ultimately the finished network has to be practically attainable, so that there is no point in optimizing on paper beyond the stability limits imposed by component tolerances.

Design Considerations

It is not necessary to have accurately linear windings for the potentiometer, although, of course, this greatly helps the preliminary analysis. Simple resistance bridge measurements can be made of the winding resistances between tapping points. Similarly the tapping points need not necessarily be equally spaced as again a simple dimensional measurement will provide values for n_1, n_2, etc.

The potentiometer's total winding resistance and the values of the shunt resistors must be considered as a whole in the final design, to determine whether the signal source will be capable of supplying the load presented by the complete network when the wiper is at the input terminal, which is the worse case.

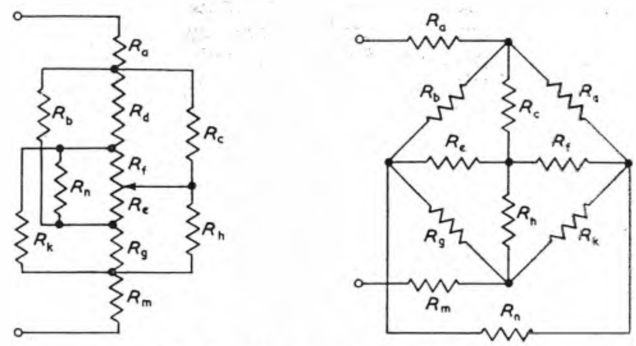

Fig. 14. Non-planar bridge network

The final design should not call for shunt resistors of excessive accuracy or stability.

Conclusions

(1) All multi-tapped potentiometers of the type described in this article can be represented by a single general network which can be further subdivided into one of four equivalent forms, according to the tapping arrangements and the position of the wiper.

(2) If equally spaced tapping points are chosen the design procedure is greatly simplified.

(3) The networks can be used to generate monotonic functions.

(4) The effect of output loading can in many cases be employed to improve the accuracy of fit of the generated curve.

(5) In conjunction with appropriate summing amplifier networks, multi-tapped potentiometers can be used to generate a wide variety of non-linear laws, e.g. secant, tangent, sine, cosine, logarithmic, etc.

Further investigation is needed to completely generalize all possible arrangements of multi-tapped potentiometers, which in their most general form become the non-planar bridge network of Fig. 14; and it is doubtful if the generalized equation for this would be simple enough for practical application.

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ANALYSIS OF IMPACT NOISE

By F. M. Savage*

Instruments for the measurement and analysis of steady noise have made it possible to establish criteria for avoiding hearing loss in personnel exposed to continuous noise. The development of impact noise analysers should provide data for dealing with the even more dangerous impact noises. In this article a small portable instrument for the measurement of impulsive noise is described. The instrument may also be used for other purposes, such as the measurement of peak accelerations when used with a suitable accelerometer.

(Voir page 247 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 253)

IT is now generally agreed that hearing loss due to noise is a function both of length of exposure and of noise intensity. Although hard and fast rules regarding the time-dependence of ear injury cannot be made because of wide variations in the characteristics of industrial noises and in human response, modern electronic equipment coupled with an intelligent use of statistical methods has provided a scientific basis for determining safe noise levels for steady noise. The classic report[1] by a sub-committee of the American Standards Association, for instance, enables a realistic prediction to be made of the average hearing loss to be expected under a given set of steady conditions.

Impact Noise

An important limitation set on this work is imposed by the word 'steady'. With conventional sound level meters and analysers it is impossible to obtain an accurate reading on any sound which does not remain steady for a period approaching the time-constant of the meter (response time of the normal measuring circuit is about 0·25sec). While the meter needle will give a brief flick in response to such a sound, it will inevitably give a maximum reading between 15 and 30dB down on the signal peak. Yet all work indicates that such impact noises are a considerably greater menace to hearing than steady noise. Quantitative estimates of this physiological difference are however still very sparse, due to the difficulty in obtaining reasonable readings of impact noise under workshop conditions. Some work, it is true, has been carried out using cathode-ray oscillographs and cameras. The cumbersome nature of this equipment however precludes its use for any widespread survey such as would be necessary to place the results on a reasonable statistical basis.

A further disadvantage in the use of oscillographs is that while a great amount of information is obtained on such matters as waveform, the oscillogram yields no direct rating of noise, which is necessary as a basis for corrective action. Such quantitative information could readily be obtained if it were possible to extend the range of the conventional sound level meter to cover impact noise.

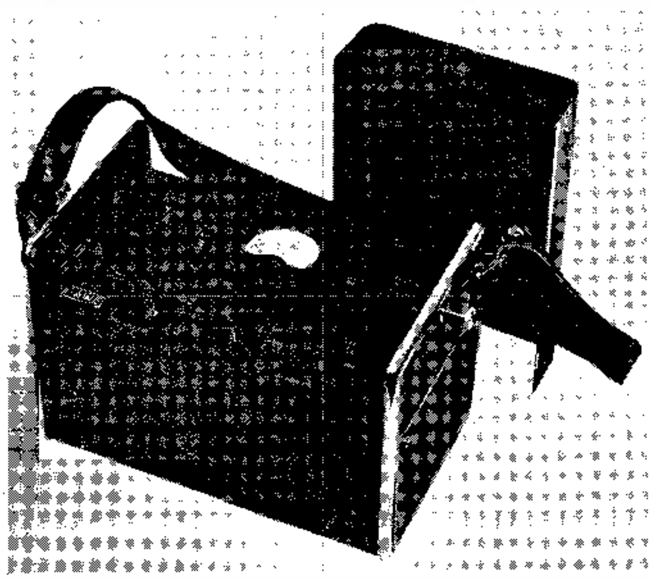

Fig. 1. The Dawe Instruments type 1412 impact noise analyser

A major problem facing the designer of an instrument for measuring impact noise is the selection of the most appropriate parameters for defining such a noise. Some guidance is available from the work of Rosenblith[2], who showed that the response time of the muscles in the middle ear to an impulse noise may be in the region of 10msec, or of the same order as the duration of the noise. This suggests that the initial peak of the impact tends to have a greater effect on the ear than the tail, and that the peak intensity may therefore prove to be the most important measure of the effect of impact noise on the human ear. In the case of the instrument described below it was therefore decided to consider peak intensity level and duration as the vital parameters, a further measurement of 'quasi peak' intensity being incorporated both for handling repeated impulses and to provide a practical method for calibration.

The peak intensity measured is the maximum sound-pressure level reached by the noise after the analyser control has been switched from the 'reset' position. The time taken to register the peak level—held by the instrument for a number of seconds for subsequent recording—is about 50µsec. Comparisons of the peak levels thus recorded have in general agreed to within 1dB with those obtained from a cathode-ray oscillograph. This means in effect that on this setting the instrument can be considered as instantaneous when operating on sound waves.

By contrast, the quasi peak reading provides a continuous indication of the higher sound pressure levels which have occurred between resetting and the time of indication. The relevant circuit has a fast rise-time below 1msec and a slower decay time of about 600msec, so that peak values can be followed reasonably well. This circuit is useful when impacts follow each other so rapidly that there is no time to take the peak reading before the next impact intervenes.

By applying a known sine wave input to the quasi peak circuit it is simple to achieve precise calibration of the instrument. On this circuit incidentally the instrument behaves substantially in accordance with the standard specifications[3] for a meter to measure electrical impulsive noise.

* *Dawe Instruments Ltd.*

The other important parameter, impulse duration, is determined indirectly. The 'time average level' of the noise is measured over a known period and the impulse duration estimated by comparing the time average and peak values. Naturally the waveform of the particular impulse will control the exact shape of the calibration curve used to derive peak duration. As described below, however, it is possible to make certain assumptions which will permit a reasonable estimation to be obtained without measuring the form factor in every case.

An important feature of the instrument is that the capacitors for measuring peak intensity, quasi peak intensity, and impact duration are permanently connected in circuit. All three capacitors are therefore automatically charged on each occasion without the need for action by the operator. Low leakage currents are achieved in the design: charges are maintained within 1dB of the initial value (for 10sec on the peak intensity capacitor and for 1min on the time average capacitor) under all normal conditions of temperature and humidity. Accurate readings of both intensity and duration can thus be obtained on the same impact merely by turning the selector switch. This feature eliminates the uncertainty caused by having to measure these two interrelated parameters on successive impacts which may well vary appreciably in magnitude and waveform.

Design of Instrument

The type 1412 impact noise analyser has been developed to provide an attachment enabling the Dawe type 1400 sound level meter to measure both impulsive and steady sounds. In addition to this prime function, the type 1412 analyser may also be used to measure peak accelerations when used with a suitable accelerometer or to analyse any impulsive waveform capable of supplying about 1V peak across 50kΩ.

The design adopted results in a readily portable instrument. Including dry batteries, the analyser weighs only 8¾lb and measures 11¼in by 7in by 8in overall. A general external view is given in Fig. 1.

Circuit Description

The instrument is completely self-contained, h.t. and l.t. power being provided by dry batteries. With the exception of the cathode-follower, the valves are sub-miniature types with a low filament current and thus a relatively low mutual conductance. As a result the circuit design required special care to provide a wide-band frequency response.

The signal is fed to a simple input attenuator RV_1 (see Fig. 2). Either the positive or negative-going pulse (increasing or decreasing pressure wave) can be selected for measurement. Negative-going pulses are switched directly to V_2 via C_7. Positive-going signals are passed to the anode-follower (V_1). The circuit constants of this stage are designed to give a gain of -1. This is achieved by arranging the value of R_1 to be slightly less than that of R_3, while the value of R_4 (supplying grid-leak bias to V_1) is very much greater than the combined impedance presented by R_1 and C_1 connected in series. The output from this stage (or negative-going pulses, if these are selected for measurement) is fed to a conventional pentode amplifier (V_2) via the capacitor C_7.

The output from V_2 drives V_3, a triode-connected cathode-follower. V_2 and V_3 are arranged to give a resultant gain of -10, providing a 10V peak level at full-scale deflexion of the valve-voltmeter. These two stages obtain overall feedback from R_7 and R_{11}, thus stabilizing the gain. High-frequency compensation is provided by C_6. This arrangement also provides adequate frequency response over a wide band. The bias voltage of V_3 is obtained from R_{13}, tapped into the cathode load. The voltage feedback introduced provides a lower source-impedance than could otherwise be achieved. This ensures rapid charging of the storage capacitors to the peak voltage through the silicon diodes, whose very high reverse resistance prevents the capacitors from discharging prematurely.

The silicon diodes MR_1, MR_2, and MR_3 are permanently connected to the output of V_3 and the pulses from V_3 are applied simultaneously to all three rectifier circuits (through resistors selected by the time-constant range switch in the case of the 'time average' circuit). Only positive-going pulses are thus offered to the valve-voltmeter. The low-leakage properties of the storage capacitors C_{13}, C_{14} and C_{15}, which are of course also charged simultaneously, provide sufficient time for the 'characteristic' selector to be switched to each of the three circuits and for readings to be taken, so that an impulse can be interpreted with all three rectifier characteristics. In the 'peak' network, C_{14} is directly charged, the peak voltage being also directly applied to the grid of the electrometer valve V_4. In the 'quasi-peak' network, C_{15} is shunted by a very high resist-

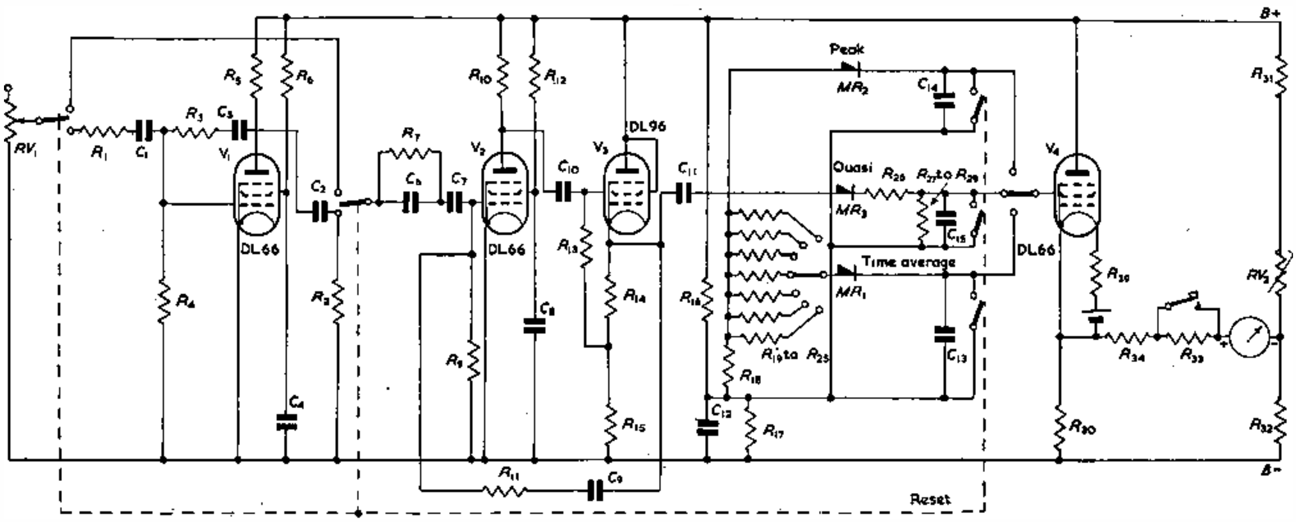

Fig. 2. Simplified circuit of impact noise analyser

ance R_{27-29} which, together with R_{26}, provides a moderate rise period with a relatively slow decay period. The characteristics of the 'peak' and 'quasi-peak' networks and the effects of varying the time-constant in the 'time-average' network are shown in Fig. 3.

Special precautions were taken in the selection of components and in the layout of the circuit in this section of the equipment to maintain the high insulation resistance.

The outputs from the three rectifier-capacitor circuits are fed to the valve-voltmeter through a selector switch. After readings are taken, the polarity switch is returned to 'reset'. This discharges C_{13}, C_{14} and C_{15}, so that a new reading can be taken.

The silicon junction diodes used in this equipment have an extremely high reverse resistance, which is greater than $10^{10}\Omega$ in the case of MR_1 and MR_3, while MR_2 has a reverse resistance greater than $10^{8}\Omega$; the forward resistance of the rectifiers is relatively low.

A simple bridge circuit is used for the valve-voltmeter. The triode-connected valve (V_4) is aged for 24 hours to secure a low grid current. The filament series resistor R_{39} ensures that this is maintained in service. Zero setting of the valve-voltmeter is effected by the potentiometer RV_2. The correct bias voltage is obtained from a divider network consisting of R_{16} and R_{17}, the valve grid itself being open-circuit.

The h.t. and l.t. voltages can also be switched to the meter to indicate the state of the batteries under operating conditions.

Estimating Impact Duration

As mentioned before, the duration of an impact noise can be estimated from a knowledge of the peak and time average noise levels and of the waveform. It may be convenient to illustrate the method employed by reference to a simple rectangular pulse.

If the time-constant of the rectifier circuit is set to t seconds and a constant voltage applied, the resultant capacitor voltage will tend to build up until it eventually becomes asymptotic to the value of the applied voltage. If, for instance, the application lasts for $t/5$ seconds, the capacitor voltage will still be 15dB down on the applied voltage. After a period of $t/2$ seconds, this difference will fall to 4dB and so on. When it is remembered that the impact duration represents the only charging time for the capacitor, this duration can be obtained directly, if rather laboriously, by plotting the difference on the graph shown in Fig. 3.

Unfortunately, impact noises are not simple rectangular pulses. Readings taken on oscillographs indicate that they tend, at least as a first approximation, to decay exponentially. Using this assumption it is possible to compute a curve to give a rapid conversion to impact duration (see Fig. 4). Comparison of impact durations obtained in this way with those measured on cathode-ray oscillographs show close agreement over a wide range of industrial impact noises.

The fact that the time-constant of the time average rectifier circuit can be varied offers a simple way of checking that the correct waveform has been assumed. Clearly if the curve of Fig. 4 is suited to the particular impact, the same impact duration will be measured irrespective of the time-constant setting. Bearing in mind that no two impacts are identical (the variation being particularly marked for sharply peaked waveforms such as those emitted by cartridge-operated hammers for driving studs into concrete), considerable confidence can be placed in results showing a reasonable agreement with a number of different settings of time-constant. Failure to achieve such agreement probably indicates incorrect assumptions on waveform. The actual variation in the readings will provide some guidance towards constructing a correct calibration curve which can then be checked by means of an oscillograph, if one is available.

Use with a Spectrum Analyser

From the previous discussion it will be apparent that the impact noise analyser can be used to meter the output from a spectrum analyser. It is not clear whether the measured

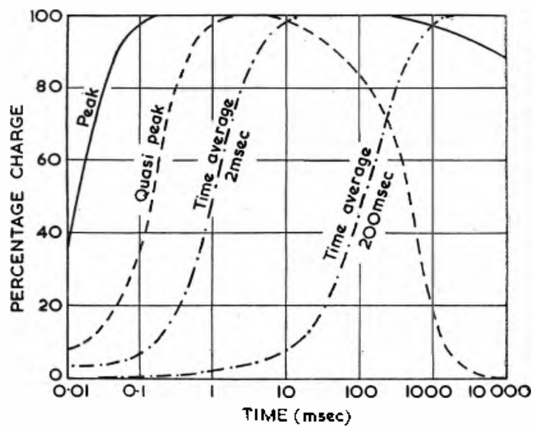

Fig. 3. Charge and decay characteristics for peak circuit, quasi peak circuit, time average circuit with low time-constant, and time average circuit with high time-constant.

By comparison of these curves it is possible to determine the decay time of any square waveform if the dB difference between peak level and averaged level is known.

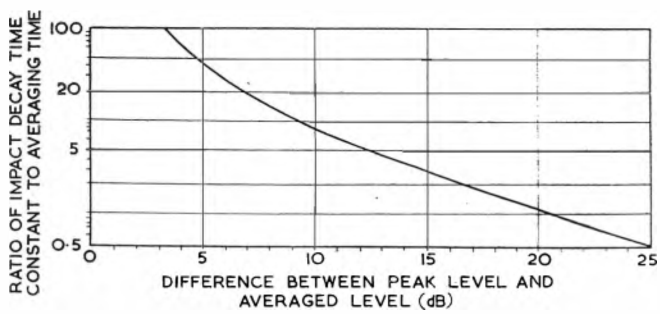

Fig. 4. Curve for calculating the duration time of exponentially decaying pulses

peak level in any band is physically significant in itself, since such a peak is not actually in the impact noise itself. In the absence of any better data, however, the reading would appear to give some numerical indication of the noise level within the particular band. It is also possible to use this peak level and the time-averaged level to determine an equivalent decay time for the noise in each band.

Having obtained such a decay time, it is important to consider the effect of filter response time. If a single tuned circuit is used as a filter, the response times will be fairly rapid. Such a filter is, however, not sufficiently selective for noise analysis. The use of more complex networks is therefore necessary to obtain greater selectivity, although this involves a corresponding lengthening of response time.

To put the whole matter in perspective, the ratio of the octave response times for a small press to those of the Dawe type 1410 octave band analyser, which is normally recommended for this purpose, never falls below about 6·6. For this class of noise at least, no special precautions need therefore be taken, since any ratio above 3:1 means that the response time of the filter is negligible in relation to the duration of the impact noise. Naturally, before this assumption is made on any new application, it is necessary to carry out measurements of the particular noise under investigation and to compare them with the curve for the design of spectrum analyser in use at the time.

Selection of Microphones

The main requirement of a suitable microphone for an impact noise analyser is that its response should be similar to that of the human ear. Since this criterion was a major factor in the initial selection of the microphone for the type 1400 sound level meter, it is not surprising that studies using this type of microphone with an oscillograph indicate that a satisfactory performance can be expected.

An interesting point arises in deciding when a particular reading on the impact noise analyser represents a compression or a rarefaction stage of the sound wave. A simple method of settling this point for the first time consisted in placing the microphone in a container and providing a definite compression by forcing air into it with a tyre pump.

Conclusion

This impact noise analyser provides a theoretically correct measurement of peak level which agrees very closely with that given by laboratory equipment. This peak level is the parameter which is required for a great deal of work on the noise reduction of such equipment as punches and forging presses, guillotines, and even guns and small arms. It thus provides the best and perhaps the only way of comparing the effectiveness of different modifications.

Measurement of peak level is also vital in any programme aimed at preventing hearing loss in operators exposed to impact noise. The success of such programmes will undoubtedly throw more light on the response of the human ear to impulsive noises and may well advance our knowledge in this branch of otology to its present level for steady noise.

An untapped field is the possibility of using impact noise analysis as a measure of the efficiency of many industrial processes. Here again an accurate measure of peak noise would seem to be an essential prerequisite for success.

In many forms of acoustic survey, the need frequently arises for measurement of the vibration originating from impacts. This demand can also be catered for merely by changing the form of transducer used with the impact noise analyser.

This facility offers considerable advantages in the investigation of such problems as 'tyre thump' which has proved so troublesome to tyre manufacturers during recent years.

The derivation of impact duration by an indirect method naturally introduces the possibility of errors not present in a direct measurement of peak noise level. This parameter is however generally required to a lesser degree of precision than is a measurement of noise level. Accordingly the accuracy obtainable appears to be more than adequate for present requirements.

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Closed Circuit Television Equipment for Advertising Agencies

Two closed circuit television and telecine equipments have recently been installed in offices of advertising agencies in London.

The first of these was installed at the headquarters of London Press Exchange Television Ltd by Marconi's Wireless Telegraph Co. Ltd. Forty-three rooms, including the Board Room of the L.P.E. building have been wired to make use of the facilities afforded by the new system.

A feature of this installation is its extreme flexibility. Two Marconi BD871 cameras have been used and each can have a dual function. Mounted on a tripod stand and fitted with a zoom lens either can be employed for viewing auditions, casting or rehearsals in the theatre. Mounted on a pedestal in the projection room either is available for telecine work. In this position they can be swivelled to 'view' the film running through either of three projectors (two 35mm and one 16mm) or film strip in a slide projector.

For telecine operation the camera scans the film frame and produces video impulses which are fed together with sound signals into radio-frequency modulators and amplifiers. These are crystal-controlled miniature transmitters. The modulators feed their outputs into cables to the forty-three receiving points. In the same way the ouput from the auditioning camera is fed to another radio-frequency modulator and amplifier and similarly distributed. The receivers therefore have four radio-frequency signals fed to them, namely the BBC and ITA signals and the signals emanating from the cameras. It is planned to use one camera permanently for telecine and the other for use in the viewing theatre or for telecine as required.

The 35mm projectors can run either 'married' or 'unmarried' prints and are modified to use magnetic stripe film. These projectors and the associated sound systems have been supplied by R.C.A. (Great Britain) Ltd, while the inter-office cabling has been carried out by Belling & Lee Ltd, both under sub-contract to Marconi's.

The second installation is that at the London headquarters of the J. Walter Thompson advertising agency.

This equipment, installed and manufactured by Pye Ltd, transmits pictures to a network of receiving screens throughout the building. The telecine installation caters for 16mm and 35mm film or slides, and has facilities for running optical or magnetic sound tracks, including a 35mm Westrex magnetic machine.

A closed-circuit 'live' television system is also available.

Estimation of Depth of Blemishes in Glass*

Glass used for optical purposes often contains small imperfections which it may be possible to eliminate in subsequent grinding and polishing. In order to know if a particular sample is usable, one must know the depth of any blemishes below the surface. This can readily be done with the following apparatus.

It is assumed that one surface is optically clear. Immediately above the blemish to be examined, a microscope, having a short depth of focus, is focused on to the clear surface, most conveniently by using a fixed stop to contact the surface, or alternatively by marking it with dye or other material.

A gauge consisting of a graduated echelon or wedge of glass (or other transparent material) is placed over the blemish and moved across until the thickness of the gauge between microscope objective and blemish is such that the blemish is brought into clear focus at which point the depth of the blemish may be read off on the gauge.

If the glass being examined has the same refractive index as the gauge, the depth is given by

$d = (\mu - 1)t$ where d = depth of blemish
t = thickness of gauge
μ = refractive index.

So that for $\mu = 1{\cdot}5$ a depth of blemish d is measured using a gauge twice this thickness, thus increasing the accuracy of measurement. Also there are no errors due to backlash in the microscope movement as when using a travelling microscope.

If the refractive indices of gauge and glass examined are appreciably different, the gauge must be calibrated accordingly, in which case

$d = (\mu_2/\mu_1)(\mu_1 - 1)t$ where μ_1 = refractive index of gauge
μ_2 = refractive of glass being examined.

This method of depth measurement is particularly suitable if it is required to work between fixed limits.

* *Communication from E.M.I. Ltd.*

The Recording and Collocation of Waveforms (Part 2)

By R. J. D. Reeves*

(Voir page 184 pour le résumé en français:
Zusammenfassung in deutscher Sprache auf Seite 189)

A Modern Strobograph

The Ekco Strobograph (illustrated on right) will be the first strobograph to be commercially available since Hospitaliers 'Ondographe' ceased production about twenty-five years ago, and the re-introduction of this technique closes the gap in present-day instrumentation by providing, once again, for the permanent recording of waveforms, but in this case waveforms of incomparably higher speeds. The significant advance that this represents over c.r.t. photography is not simply that of functional simplicity but that the presentation is precisely controlled in amplitude, so that measurements can be taken from the strobogram with a ruler at any subsequent time. Equally important is the fact that the instrument is direct coupled, so that multiple recordings can be made showing the spacing between related waveforms on the voltage scale with an accuracy no less than that provided for the measurement of the individual waveforms. This feature is not necessarily peculiar to strobographs, but is so rarely available that it provides a quite unaccustomed luxury to be able, for example, to collect the complete set of valve electrode waveforms properly disposed on the voltage scale for the true assessment of the instantaneous working conditions.

A more complex form of this instrument has two strobing channels and provides for the recording of families of curves being explored under periodic or pulse conditions, examples readily coming to mind being pulsed valve characteristics or high speed *B-H* curves.

Typically, these two applications depend upon the use of auxiliary apparatus specially designed to explore the characteristics in a recurrent manner, and it is envisaged that the two-channel strobograph will be mostly employed in a variety of specialized applications rather than the analysis of functional circuits. At 'mechanical' frequencies this form reduces to a simple *X-Y* plotting table with virtually continuous tracking.

Both forms of the machine are Class 2 with a vertical scale accuracy better than ½ per cent. An electronic sampling pulse provides a time resolution better than 3mμsec, and a voltage controlled delay system has a linearity better than 5 per cent. The servo system is a velocity lag type using a 50c/s a.c. motor and tachogenerator, and the strobe pulse is manually advanced by a handwheel.

There are three scale ranges per decade, covering sensitivity ranges from 50V/in to 0·5V/in, and time scales from 500μsec/in to 0·1μsec/in, the recording paper being 10in by 7½in.

The repetition rate generator is variable from 200c/s to 15kc/s and can be locked to repetition rates up to 150kc/s, with aural and visual indication of the synchronizing condition.

Mechanical Features

The instrument is built into a mobile rack frame, with side doors giving ready access to the units. Each unit has vertical chassis with components on the outside and valves projecting inwards. This arrangement permits a planned airflow in the centre of the unit, using wide apertures beneath and above the doors. At the very bottom of the frame a drawer is provided for recording paper.

The plotting table is tilted up at 45° to provide perpendicular presentation to an operator standing in front of the machine, and this arrangement conveniently submerges the pen drive mechanism beneath the main table. In this way, a clean appearance is obtained and there is ample space on the table for the annotation of strobograms.

* *Solartron Electronic Group Ltd, formerly E. K. Cole Ltd.*

The above photograph shows the Ekco strobograph

The recording pen is actuated by a steel tape drive which differs from usual in the use of bowed tape in place of flat tape, which can transmit thrust as well as tension forces and therefore requires no return loop. This tape is fed off a flat drum and takes up the bowed form as it leaves; the thrust being provided by a second 'backing' tape which is fed off with it and is reeled up on a drum on the shaft of the reference potentiometer.

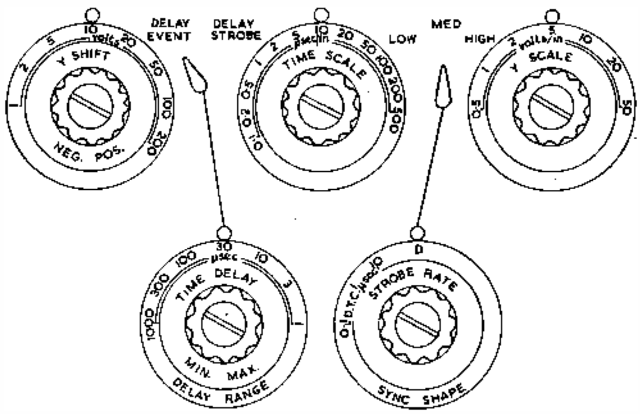


Fig. 10. Control panel

The illustration on page 204 is of a double channel machine with two signal probes. Notice the small complement of controls for this very general purpose instrument, and in the case of a single channel machine the controls number two less, namely five, which is illustrated in Fig. 10. This simple layout results from the integration of several related controls into one knob, and from the fact that in a strobograph there is no focus, brightness, astigmatism, set zero etc., and in fact the only incidental control is a loudspeaker on-off switch.

In each case where a control covers a wide range (namely those concerned with 'shift' and 'repetition rate') the range switch is operated by the limit stops of the fine control, and is actuated in a natural way at the limit of the fine range. A third control can be engaged by a dog clutch brought into operation by pressing the knob. Where this is incorporated a separate flag is necessary to indicate the setting of the third control.

A shaft which is identified with table movement is available at the rear of the instrument for the attachment of an external potentiometer or synchro, so that the X scale can be identified with some parameter other than time. By this means using the Y channel as a recording d.c. voltmeter, the instrument may be employed to plot families of response curves, characteristics etc., without recourse to two-channel working.

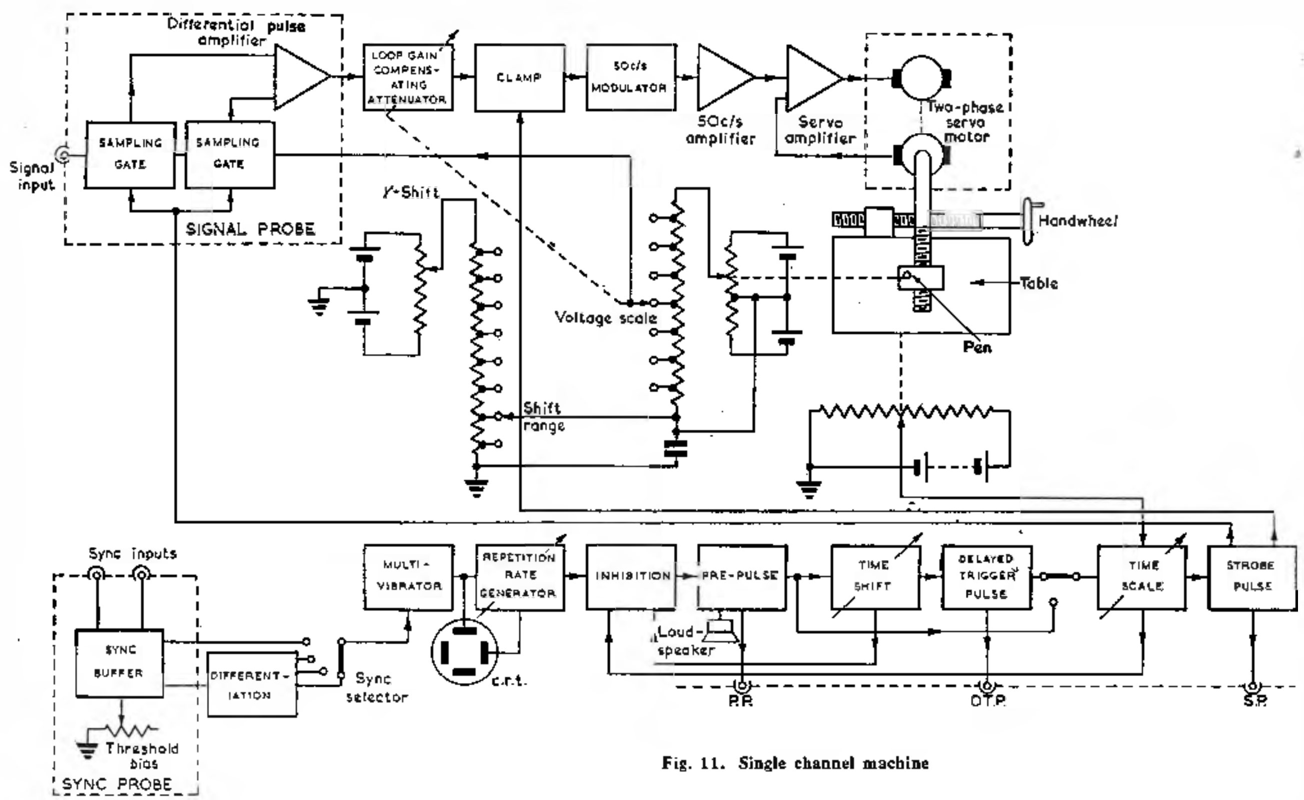


Fig. 11. Single channel machine

The drive to the moving table is a simple rack and pinion arrangement with anti-backlash coupling to the pick-off potentiometer. In the double-channel instrument where the table has to track a signal waveform a specially light table is constructed of aluminium honeycomb.

Scales are provided on the pen and table not as a part of the measuring procedure but as an aid to operating, so that significant instants or voltages can be noticed. A separate scale is essential, however, for the handwheel if this is not associated with X displacement.

Fig. 11 is a block schematic of the single-channel machine and the function of the component units is described in the following sections.

The 'Repetition Rate' Generator and Synchronizing Arrangements

A probe is provided for the synchronizing signal so that if this should happen to be the input signal itself the minimum loading of the source is obtained. This probe is carried on a separate lead to the signal probes for greater flexibility and to avoid any possibility of crosstalk within the instrument.

One double triode is used in the probe, connected as a long-tailed pair; synchronizing signals being applied to one grid or the other depending whether the intention is to synchronize at a positive excursion in the waveform or a negative going one. This valve has a threshold bias variable from 0 to 50V which can be used to suppress interference or unwanted parts of the waveform below the indicated level, and the prepulse in the instrument will be triggered as it crosses this level.

The output is connected to a slave trigger circuit which produces a square waveform of fixed amplitude which follows the synchronizing signal as it crosses and recrosses the synchronizing threshold. It is completely absent if there is no synchronizing signal. This waveform is used to synchronize the repetition rate generator and is also displayed on a small c.r.t. as a visual aid to synchronizing.

The 'repetition-rate' generator is a free running relaxation oscillator, continuously variable in three ranges from 200c/s to 15kc/s. This oscillator takes the form of a Miller transitron circuit with a controlled bottoming level and auxiliary flash-back valve. The fine tuning is by control of the grid potential and the range switching by the Miller capacitor. A linear stroke circuit of this kind is to be pre-

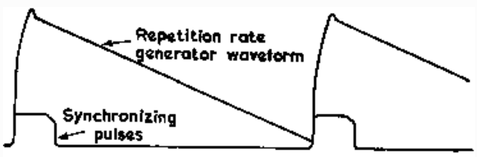


Fig. 12. Synchronizing the repetition-rate generator

ferred to the simpler exponential wave generator for two reasons, one being simply that the waveform can be used as a display base in the oscilloscope, but more important is that the instant of flash-back is well defined, being the time at which the run down traverses the bottoming potential. This is less prone to jitter than an exponential approach.

The (shaped) synchronizing wave is superimposed on the bottoming potential and the positive excursions momentarily raise the bottoming level and thus shorten the run-down period if the run-down is near to terminating at that time. In this way the synchronizing signal has a tendency to raise the repetition frequency and lock it into its own rate, or a sub-harmonic thereof, the condition being obtained by suitable adjustment of the free rate control. The repetition-rate generator waveform and typical synchronizing pulses are shown on common time and voltage scales in Fig. 12.

Sub-harmonic working is achieved by limiting the amplitude of the trigger waveform so that only trigger pulses occurring near the natural bottoming time are effective and earlier ones are missed, the simplest case being illustrated in Fig. 13. A small amplitude trigger pulse makes a large division possible but requires accurate setting of the rate control. A large trigger pulse on the other hand, reduces the division ratio that is possible but is less critical of the free rate setting. Reliable division up to 10:1 is a suitable compromise to aim for.

The repetition-rate generator in its turn fires a blocking oscillator, the pulse from which is termed the 'pre-pulse' and initiates the time delay circuits. This happens when the repetition generator flashes back, and so immediately follows the incoming synchronizing signal. The pre-pulse is also brought out to the front panel for use as a marker or trigger pulse.

Interposed between the repetition-rate generator and the pre-pulse generator is an inhibitor circuit which prevents the pre-pulse being fired if the delay circuits have not completed the previous strokes, which could cause faulty operation of the delay circuits. By virtue of the inhibitor the pre-pulse frequency may be sub-harmonic to the repetition-rate generator just as, in its turn, the repetition rate generator may be sub-harmonic to the synchronizing signal.

A small loudspeaker is connected to the pre-pulse generator (or the subsequent strobe pulse generator) and provides an aural aid to synchronizing, this completing the features available for synchronizing shaping and monitoring.

In the absence of a synchronizing signal the repetition-rate sounds pure at all frequencies and can be tuned in a continuous manner. In the presence of a synchronizing signal the repetition-rate frequency proceeds in a series of jumps as it is tuned past the sub-harmonic frequencies, and pulls in strongly to the fundamental. At frequencies between the sub-harmonics or above the fundamental the tone sounds harsh as some repetition periods are shortened

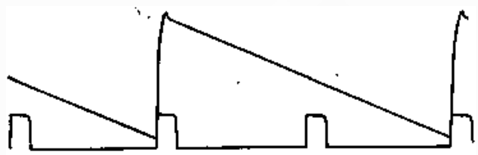


Fig. 13. Sub-harmonic working

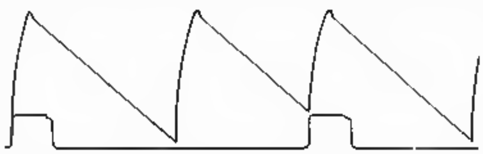


Fig. 14. Harmonic locking (incorrect)

but not all. By this means the proper settings are easily distinguished and even one 'slipped' pulse is easily heard. An incorrect setting of the repetition-rate generator is illustrated in Fig. 14 which produces pulses at twice the signal repetition rate which is an inadmissable condition for the purpose of strobing.

On the c.r.t. the synchronizing signals are displayed with the repetition period as a display base and the significant synchronizing pulse which actually triggers the repetition period may or may not be visible depending on its duration. But the missed triggers are all visible and if these are stationary and tend to remain so in spite of small adjustments of the rate control then the instrument is firmly locked, and the number of visible pulses reveals what sub-harmonic is being employed. Faulty tuning below the fundamental frequency will show a blurred display or unevenly spaced pulses which are not stable in position.

The functioning of the inhibitor is not indicated on the c.r.t. but is apparent on the loudspeaker, which always indicates true strobe rate. Switching to time scales which are longer than the repetition-rate generator period produces sub-harmonic tones.

Careful measurement reveals that when pull-in is apparently accomplished there is residual jitter between the input synchronizing signal and the pre-pulse which is proportional to the amount of pulling which is necessary to enforce the synchronism. This is not indicated on the monitors, so when the fastest time scales are being used a mental note must be made to tune to the high frequency end of the locked arc on the rate control, where the free rate is most nearly equal to the synchronized rate.

It may appear that these arrangements for synchronizing are rather elaborate, but this complexity arises partly from a desire to provide comprehensive facilities in the instrument and partly by the special problems of synchronizing a strobograph. As the waveform presentation is not mobile there is no direct indication of the synchronizing condition unless special aids are provided, and so some auxiliary presentation is essential. Furthermore, it is not permissible to use a triggered repetition-rate generator because the motor servos depend on the recurrence of the strobe at a minimum repetition-rate in order to keep the control loop closed. This is why the repetition-rate generator is designed to free-run in the absence of a trigger signal and is properly described as a synchronized oscillator rather than a triggered one.

Clearly, successful strobing depends upon accurate synchronizing and in a general purpose instrument every effort must be made to provide this over a wide range of conditions.

Time Delay Circuits

There are two independent delay circuits, one being associated with time scale and controlled by the handwheel, and the other with the time shift control. The circuits are similar but the time shift circuit is not accurately calibrated and has only two ranges per decade.

A linear run-down and flash-back is produced in just the same way as in the repetition-rate generator, but instead of free running these circuits are of the one-shot type, and wait at the start potential until a trigger pulse is received. This is accomplished by returning the suppressor grid leak of the Miller valve to a cut-off bias and using the trigger pulse to overcome this bias and initiate the run-down, the action being sustained by regenerative (transitron) coupling between screen and suppressor. Flash-back is automatic when the bottoming potential is traversed.

The bottoming potential is fixed so that the run-down stroke is of fixed amplitude; but variable delay is produced by a pick-off circuit which fires a pulse when the run-down crosses a variable pick-off potential. This circuit does not react on the run-down in any way, so the paralysis time of the delay circuit is independent of the pick-off potential and depends only on the delay range. This in an important point when the range called for exceeds the repetition period of the synchronizing signal because then, through the medium of the pre-pulse inhibitor, the time delay circuits govern the strobe rate, and this must be independent of the strobe position or the synchronizing setting could change in the course of a plot.

The time shift delay circuit is always triggered by the pre-pulse, and the pick-off circuit associated with it fires a blocking oscillator termed the delayed trigger pulse generator. The delayed trigger pulse can therefore be made to occur at any time subsequent to the pre-pulse according to the settings of the time shift and shift range controls.

The second delay circuit, the 'time scale' delay, may be initiated by the delayed trigger pulse so that the time scale represented on the paper is delayed with respect to the pre-pulse. This mode of working is called the 'delayed strobe' condition and it corresponds to normal X shift facilities in other oscillographs, but an alternative mode of operation is provided called the 'delayed event' condition, in which both delay circuits are initiated by the pre-pulse and the delayed trigger pulse is fed out to trigger the circuit under observation. When the external apparatus can be 'slaved' to the strobograph in this way the X shift control is apparently reversed and the strobe is advanced with respect to the event by the shift control, and the range of the presentation is extended to times prior to the event under observation instead of being confined to subsequent times, as would be the case in the delayed strobe mode if there were no prior synchronizing signal.

This mode of operation is not so restricted in application as might be imagined, because the strobograph can be freely set to a repetition-rate appropriate to the external circuit, and experimental pulse circuits may be deliberately constructed without a repetition-rate generator with the intention of slaving them to the strobograph.

The time scale is normally the X scale of the paper, but when two channel working is employed, the time scale is parametric and follows the convolutions of the trace. In this condition a separate scale coupled to the handwheel indicates the progress of the strobe pulse and the trace can be annotated from this if required.

In the case of the time scale circuit the pick-off potential is controlled by a precision potentiometer energized by a stabilized power supply, and rigidly coupled to the X co-ordinate of the table (when appropriate) so that the

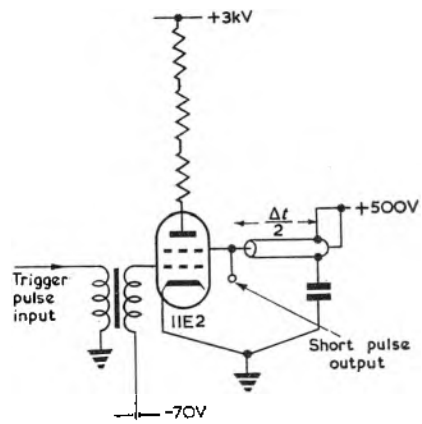


Fig. 15. 'Screen step' circuit

accuracy of the time scale is not appreciably less than that of the waveform of the delay circuit. The linearity of this may be considered to be slightly greater than is customary in a cathode-ray oscillograph because it is not paraphased or loaded for any purpose other than the pick-off circuit. The pick-off circuit itself is isolated from the Miller waveform by a diode, and the onset of conduction here initiates the output pulse so that any loading this circuit may produce is subsequent to the pick-off and not relevant to the linearity. This pick-up action has some similarity to a multiar but uses a two-valve trigger circuit with a slave blocking oscillator to give a firm indication of pick-off.

The delay range switching is by the Miller capacitor which is accurate to ± 1 per cent and the variation in bias on this valve may cause the discharge current to vary a further $\pm \frac{1}{2}$ per cent, so that altogether it will be possible to state the time scale (μsec/in) to a precision better than 2 or 3 per cent, but it can be more accurately calibrated by supplying specimen strobograms on all scales of precision frequency oscillators.

Thus without employing the special methods peculiar to strobography for controlling the time scale the accuracy may still be comparable to the best that is currently available.

Strobe Gate Generator

The pick-off circuit associated with the time scale delay fires a pulse called the strobe marker pulse, which has three functions to perform. It has to trigger the actual strobe gate generator which is the ultimate function of the time delay circuits, it has to drive the demodulating clamp which converts the error pulse from the signal probe to a d.c. signal and it is fed out for external use to mark the strobe instant if this should be of some interest in the procedure.

The strobe gate generator is the very heart of the instrument and might well have assumed some complexity, but in fact it is realized in a very simple form indeed. The essential action is obtained by rapidly bottoming a tetrode or pentode which has a sharp knee in the anode characteristics, and in this way causing a heavy current to suddenly divert from the anode to the screen. The resulting screen current step is differentiated by a shorted delay line to produce the required short pulse. This 'screen step' may be brought about in two distinct ways, of which Fig. 15 shows the method which is the simplest to understand. The strobe marker pulse which itself is a moderately fast blocking oscillator pulse of 0·05μsec rise-time and 0·5μsec duration, drives the tetrode grid from cut-off to about 50V positive and finally produces about 2A of anode current. The anode load is a high value resistor so that the anode current flows into the stray capacitance and causes a rapid fall of anode potential. The speed of this descent is proportional to the current and will not reach a maximum until the drive pulse has reached full amplitude. In order to allow time for this build up the anode is returned to a very high potential so that bottoming does not occur before 0·05μsec have elapsed. Although some screen current flows from the beginning according to the usual partition laws the delayed screen step is over 1A and is easily distinguished by a differentiating line.

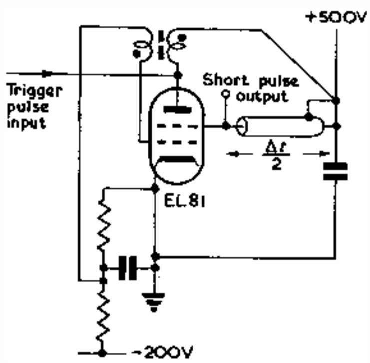


Fig. 16. Regenerative 'screen step' circuit

It will be seen that the switching action takes place entirely within the electron stream of the valve and the technique may be fairly described as an electron switch. The most apparent drawback to this arrangement is the necessity for a high anode supply voltage, although there is negligible current drain on it. For this reason the circuit of Fig. 16 was investigated and made to operate from normal supplies in an even more satisfactory manner. The intention here is to retain the screen step action but to have the valve drive itself into conduction by a regenerative transformer. This ensures that the anode descends no faster than the grid is rising without a high anode delaying potential, but the problem is now one of designing a transformer which permits a very rapid rise of voltage on both these electrodes. This is tantamount to designing an extremely fast blocking oscillator as such, and certainly the dividing line between this circuit and a conventional blocking oscillator is a fine one, but the screen connexion produces a significant difference in performance. The transformer waveform is marred by the superposition of the inevitable trigger pulse which must be applied to initiate the blocking action, and this is a comparatively slow waveform. Also the valve does not perform well when the bias is still near cut-off and it is commonly recognized that blocking oscillators are 'slow starters'. Consequently the transformer waveform only exhibits the maximum rate of rise when the step is well under way, and any tertiary output is contaminated with the slow start. On the other hand the screen output waveform has a very sudden onset at the instant of bottoming and quite apart from that the rate of rise obtained is normally better than that seen on the transformer. It must be remembered that this edge is not concurrent with that on the transformer and although the speeds of the two are related they are not manifestations of the same current avalanche but successive ones.

The best valve for this purpose discovered so far is the EL81 and the transformer used in conjunction with it is illustrated in Fig. 17. The cores are 3½cm lengths of Ferroxcube A4 material with o.d. 4mm and i.d. 2mm. It will be seen that this provides an extremely short magnetic path. The primary and secondary are each two turns of

Fig. 17. Strobe pulse transformer

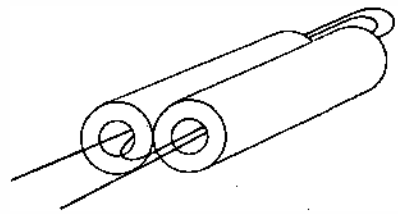
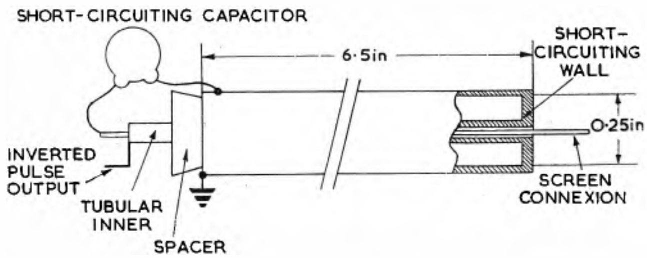


Fig. 18. Differentiating transformer



30 gauge X grade Lewmex copper wire with the shortest possible connecting leads, the top cap anode being returned via a 'skeleton' coaxial tube over the valve to minimize the inductance of the connexion

With 500V across the valve the EL81 drives itself into four or five amperes of anode current of which three to four are transferred to the screen, but in this application it is convenient to employ the normal h.t. lines and be content with lower values.

The pulse that appears across the differentiating line is necessarily of negative polarity and has to be inverted for use in the probe. This is done quite simply by providing the line with two inners in close proximity and connecting them to the outer at opposite ends. The line may then be thought of as a distributed differentiating transformer which inverts the signal. The pulse is attenuated through the device somewhat because of the capacitive loading at each end and because of the coupling factor. Coupling is maximized by having one hollow inner with the other threaded through it. The arrangement is shown in Fig. 18.

Signal Probe

The manner in which the strobe pulse is used for sampling the signal has already been outlined. The detailed arrangements are shown in Fig. 19. The two pentagrids have a cut-off bias on grid 1 and are gated on together by the strobe pulse. Grid 3 of the left-hand valve is the input terminal for the signal, and the corresponding grid of the right-hand (reference) valve is connected to the feedback potential which is associated with the Y co-ordinate of the

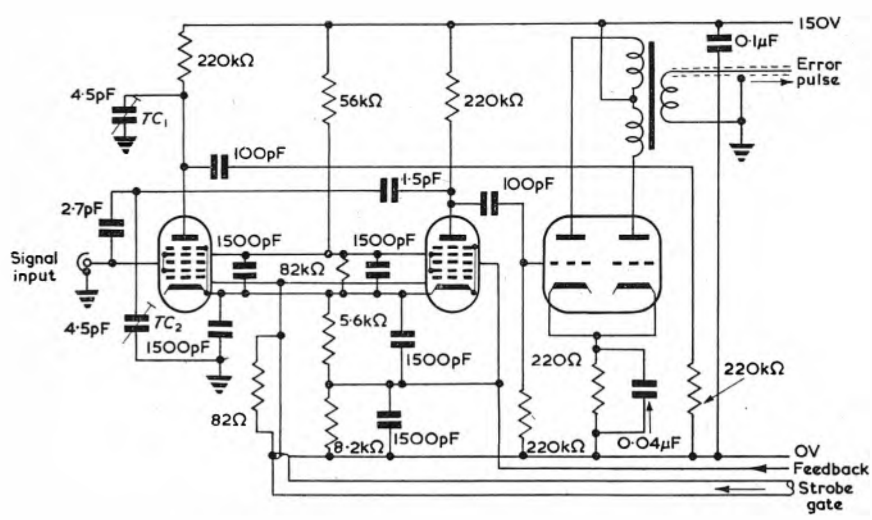


Fig. 19. Signal probe circuit

through. In the circuit diagram TC_1 is the neutralizing trimmer and TC_2 is the capacitance balance.

The Servo System

The sequence of events described so far, relating the acceptance of a synchronizing signal to the production of an error pulse is that concerned with the strobing of the input waveform, whereas the residual functions are concerned with the servo problem of tracking. Reference to Fig. 11 shows how the voltage from the slider of the pen monitor potentiometer is attenuated by the Y scale control to provide any suitable feedback coefficient of volts per inch. This control, therefore, directly alters the servo loop gain and would cause fluctuations in stiffness if it were not compensated elsewhere, but the ladder attenuator immediately after the probe offsets this effect so that the loop gain is independent of the sensitivity setting. From this point onward the error sensitivity bears a fixed relation to the pen displacement.

In order to minimize the necessary mechanical maintenance in this equipment the decision has been taken to employ a.c. servo motors in preference to d.c. so that the commutator is eliminated. The error information which is encoded in pulse amplitude has therefore to be translated into a proportional 50c/s signal, and the first step in this process is to translate it into d.c. This is accomplished by the two diode shunt modulator or 'keyed clamp' of Fig. 20, which is triggered by the strobe gate. Adjustment of

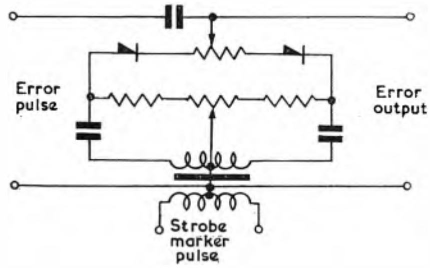
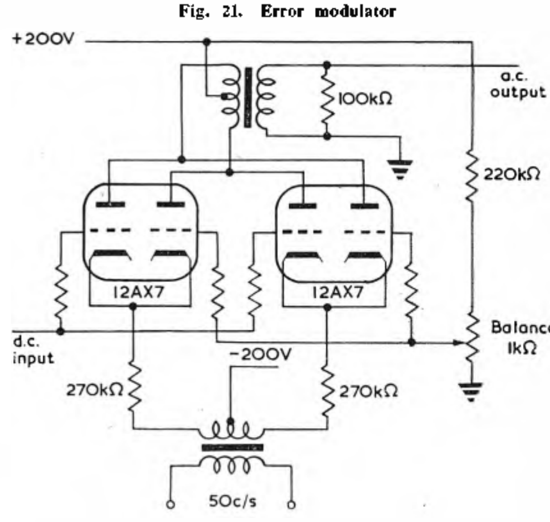


Fig. 20. Error clamp circuit

Fig. 21. Error modulator



pen. The power supplies to the probe unit are referred to this potential, not earth, so that at balance the working conditions of the input stage are constant.

As a consequence of the strobe gate each valve passes a pulse of anode current of magnitude depending on the grid 3 potential. If at that instant the signal voltage is the same as the feedback potential both valves are operating under the same conditions and the two pulses are equal in magnitude, otherwise the pulse from the input valve will be larger or smaller depending whether the signal is more positive or more negative than this. The anode resistors are large, so that the current pulses charge the stray capacitors and produce a negative voltage step with time of rise equal to the effective gate duration. This step relaxes to its former value with a time-constant determined by these stray capacitances and the anode load. The negative voltage pulses so formed are fed through a buffer amplifier to a differential pulse transformer, which feeds out, at low impedance, a pulse termed the 'error' pulse which is related in sign and magnitude to the difference between the signal and feedback potentials at the instant of strobing.

Some of the practical considerations which arise in this process are as follows: The shape of the strobe gate pulse is not rectangular and consequently the effective sampling time depends very much on the grid 1 bias, and since the anode stray capacitances integrate the valve currents over the gate period the error pulse amplitude is directly affected by variations in gate length, and this is one of the factors determining the servo loop stiffness. It is important, therefore, to stabilize the supplies to the strobe gate generator and also the bias on grid 1 of the probe valves.

The anode pulses which are to be subtracted do not have a well-defined decay time, this being governed by stray capacitances. If their decay times are not the same, differencing will never produce a complete null, and for this reason a trimmer is introduced to make provision for balancing these stray capacitances.

This procedure is integral with the process of neutralizing the circuit from the effect of capacitive injection of the signal input from grid 3 to anode. Residual capacitance from grid 3 to anode is unavoidable and its effect is serious because it enables signals to by-pass the strobe gate but its effect can be balanced out by providing a corresponding coupling from the input to the anode of the reference valve. A very small capacitance is involved here, so in order to permit practical component values the input is first attenuated by a capacitive divider before coupling

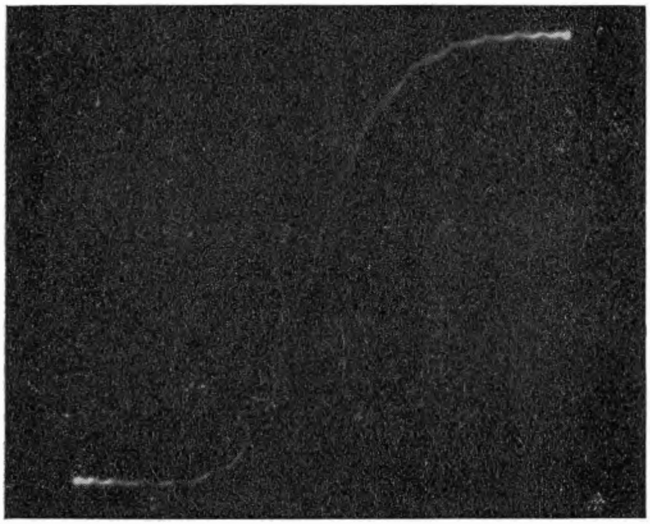


Fig. 22. Error channel transfer characteristic

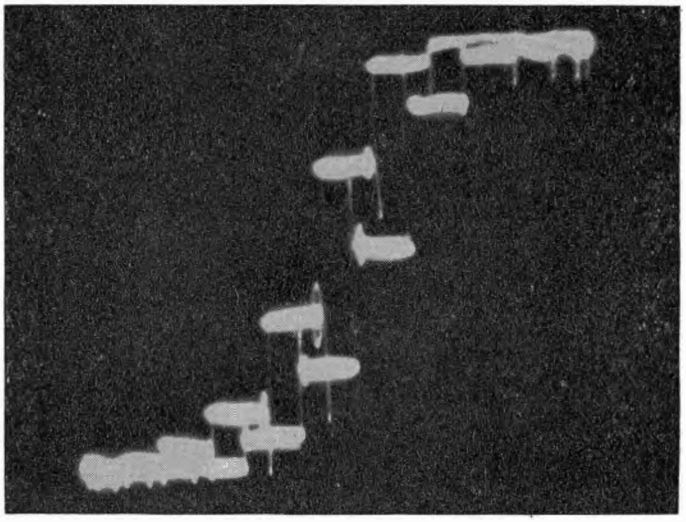


Fig. 23. Sampling lags in error channel

both the forward resistance and the relaxation time constants is necessary if it is to behave absolutely independently of pulse repetition rate.

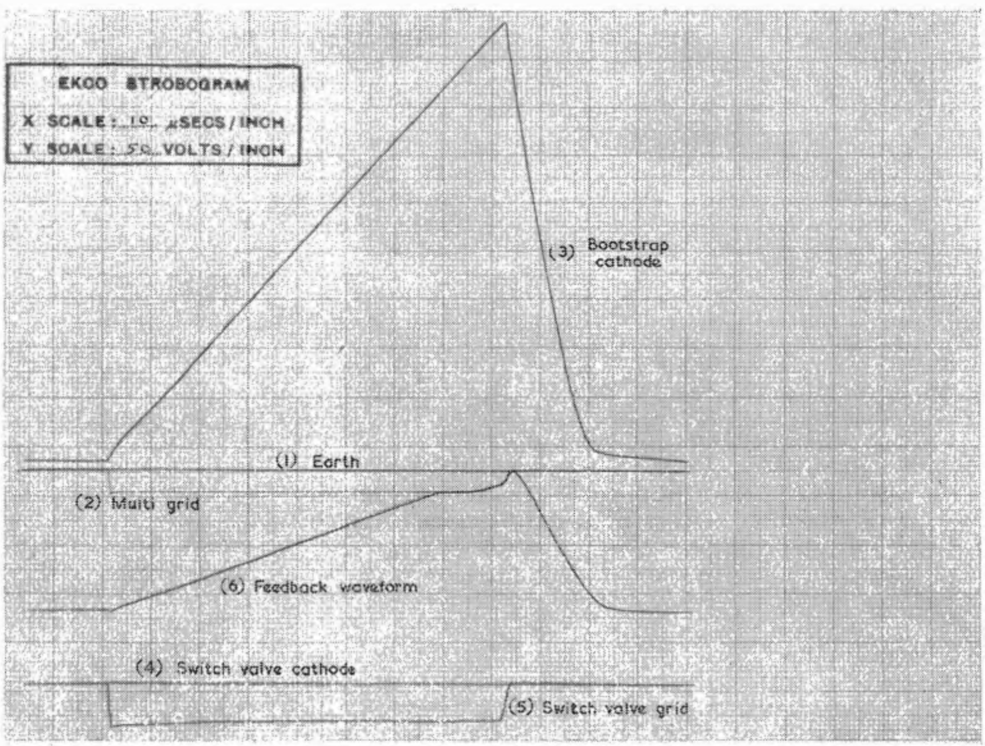
The conversion of d.c. to a.c. is a familiar problem but the restrictions associated with it in connexion with servo applications precludes many of the possibilities. In the first place any sampling process such as thyratrons or a magnetic amplifier is unacceptable because of the bandwidth limitations this imposes on 50c/s systems. The classical diode modulators are free from this defect and operate instantaneously, but since they produce square not sinusoidal waveforms these cannot be employed either. This is not because such a waveform cannot be applied to the motor itself but that at a subsequent point in the servo amplifier the error signal must be opposed to the sinusoidal tachogenerator on the motor shaft to introduce controlled damping. After this differencing point the harmonic content is increased by the tacho loop gain and will saturate the power amplifier.

The modulator that is used should be a true suppressed-carrier modulator, with undistorted output and instant response, and the choice narrows down to the use of valves as variable impedance elements. Fig. 21 shows the arrangement in which there are two similar long-tail pairs with a (common) differential transformer in the anode circuits. The modulating signal is fed to one grid of a pair and alters the distribution of the constant current from the 'tail'. Since the mutual conductance of a valve varies with the standing current this process effectively modulates the incremental impedance of the cathodes so that a superimposed 50c/s carrier current is also proportionally controlled. When the grids are at the same potential the carrier current divides equally and is annulled in the anode transformer. Variations in the relative grid potential then produce a 50c/s difference signal in this transformer. The other pair is paraphased with this to produce the same effect as far as modulation of the carrier is concerned, but having direct current surges of the opposite phase. Provided the valve characteristics are reasonably well matched these surges sum to zero in the transformer.

After adding the tacho signal a conventional power amplifier is used to drive a two-phase servo motor in which the phase shift for the reference winding is obtained by dropping a high voltage supply with a capacitor.

The fact that the error information is being sampled does not influence the behaviour of the loop provided the

Fig. 24. Strobogram of 'bootstrap' waveform



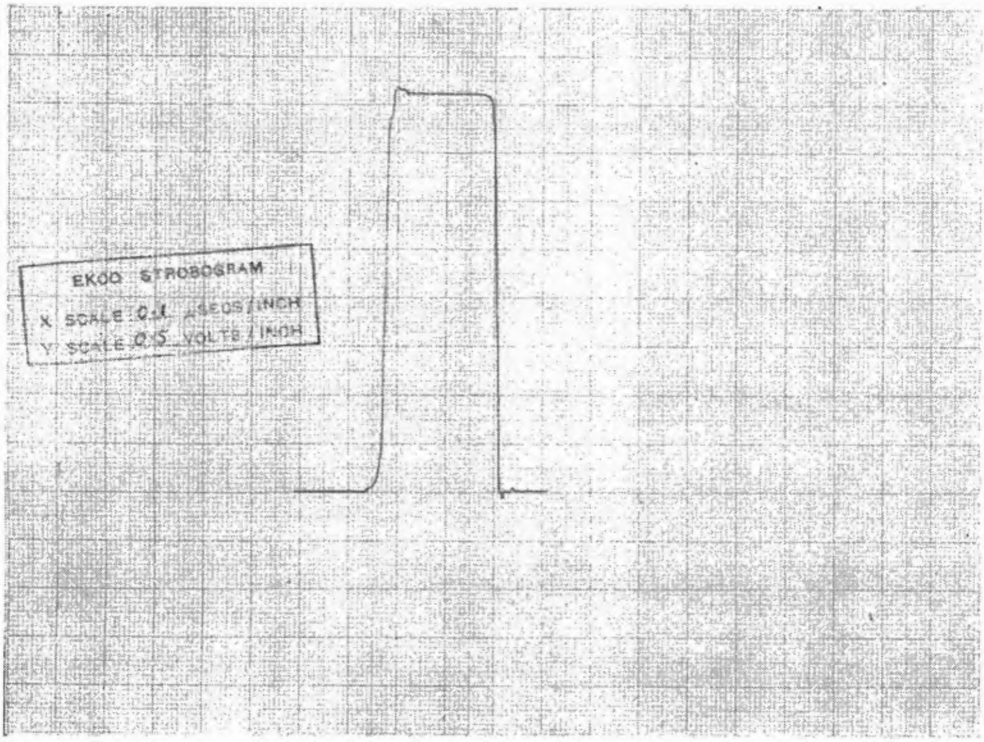


Fig. 25. Strobogram of fast pulse

shift and sensitivity settings, so that the table follows the strobe pulse as though it were mechanically linked. As a refinement the strobe position may now be picked off the table movement to eliminate tracking errors. All these operations are performed by plugging the X probe into a socket provided under the table top, in which the centre pin carries the handwheel position signal and the sleeve operates a 'metamorphosis' relay to accomplish the other changes. In this way there is no explicit control associated with the transition between the two states and there is no possibility of partial transition due to incomplete operating procedure.

The strobe source is bifurcated for the X and Y probes after the distributed inverting transformer and the gates occur simultaneously. There appears to be no occasion to require displaced gates in the two channels.

sampling rate is high compared with the bandwidth. In this case the servo bandwidth is about four cycles, but significant lags are introduced when the error is being sampled at less than 150c/s and it is this factor which limits the lowest permissible repetition-rates and time scales. This is typical of the Class 2 machine.

Fig. 22 shows the transfer characteristic of the error channel as revealed by a low frequency input excursion with the servo motor switched off. Fig. 23 shows the dispersion of this characteristic when the sampling rate is low, which, of course, is similar to a phase lag in the response.

Another characteristic of this instrument is a maximum velocity constraint due to probe saturation. The probe which is handling high level signals has a limited range of error for which it can provide proportional indication, and on the lowest sensitivity range this may represent quite a small range of pen movement. In the steady state the pen velocity is proportional to the error existing but in the presence of gross errors probe saturation will limit the pen velocity. Gross errors are typical when the input is being reconnected, and the effect is that the pen is apparently less lively than on the more sensitive ranges. In fact, within the range of acceptable error which means under typical tracking conditions, the pen behaviour is independent of the sensitivity range.

Two-Channel Working

That type of machine which provides for two-channel working must also be capable of working as a single channel system and simple arrangements provided for the transition between the two states. The duplicate channel comprises signal probe, error amplifier, servo power amplifier, and servo motor, and the handwheel is no longer mechanically coupled to the X traverse. As a two-channel system the handwheel merely shifts the strobe pulse through the time scale while both X and Y are tracking their respective input signals. To reduce to single channel working it is necessary to obtain a position signal from the handwheel and route this through the X channel with fixed

Conclusion

The machine that has been described in the second part of this article has been designed for the most general application in a field which has been formerly reserved for cathode-ray oscillography, and will no doubt find a very wide application. The time resolution and voltage measuring accuracy probably exceeds that of commercial cathode-ray oscilloscopes while the time scale accuracy and sensitivity ranges are of corresponding standard. The low frequency end of the time scale ranges is curtailed in order to retain the overall feedback feature, although this part of the spectrum is not denied to strobographs, as evidenced by the specialized machines now in use.

The extremely flexible possibilities for multiple presentation with mixed or common scales makes the method advantageous even before considering the value of the permanent recording, and the significant thing about the strobogram is that it is produced to an accurately controlled scale, so that the decision to perform a measurement or not may be deferred indefinitely. The only immediate annotation that is necessary is to indicate what the scales are.

The possibility of multiple recordings may appear to be nothing but a facile attempt to claim comprehensive facilities where none exist. In fact, the method of time sharing in one channel, even on a lethargic scale, is the best method to ensure identical scales in the collocation of the waveforms, and in this way the indicated intersection of waveforms may be relied upon absolutely.

This point is evident in the strobogram of Fig. 24, which is a family of waveforms from a 'bootstrap' time-base. These were plotted successively without adjustment of the controls so they are correctly disposed on a common vertical scale.

Waveform (1) is the earth potential from which all measurements may be taken; waveform (2) is one grid of a Schmidt trigger which is controlling the time-base; waveform (3) is the cathode of the bootstrap valve; waveforms (4) and (5) the cathode and grid respectively of the switch valve which releases the bootstrap. The bootstrap valve is seen to wait 5V above earth until the Schmidt grid goes negative (12V), which causes a simultaneous movement of the switch grid (22V). Waveform (6) is an attenuated form of the linear stroke produced by a divider chain, and is coupled to waveform (2) by a diode with the intention of terminating the time-base stroke by hauling the Schmidt grid back to earth. In fact the impedance of the divider chain is too high to accomplish this at once and the two waveforms coalesce for a short time. There is no doubt in this display what the indentation in waveform (6) is due to, whereas a display of these waveforms with indeterminate d.c. levels would merely show the deformation in (6) as a mysterious feature because there is no corresponding deformation in (2).

Fig. 25 gives a practical indication of the speed and sensitivity of the machine with a record of an exceedingly well formed $0{\cdot}1\mu$sec pulse of 2V amplitude*.

Both of these are unretouched strobograms recorded on a background of ¼in squares, Figs. 12, 13 and 14, are also from strobograms but have been traced over with a thicker pen to permit considerable reduction in size.

There is insufficient space to adequately illustrate more specimen strobograms here but further instances may be pointed out in which waveform recording is particularly valuable. For example in the design of equipment it may be desirable to estimate just what improvement a certain modification has upon the output waveform.

The superposition of records with and without the modification affords a direct comparison. Again, in a triggered train of range marker pulses it is important that the pulses from the very beginning are all the same shape, since their centroid of area is the significant parameter. Using time shift to superpose these pulses in the record permits this comparison. In cases where the area under a waveform is significant (particularly in parametric plots such as B-H curves) a strobogram permits the direct application of a planimeter. Testing waveforms to pre-drawn tolerance boundaries is also a characteristic application. In all applications the merit of the method is not simply in revealing to the operator the condition of the test circuit but in enabling him to substantiate his report with graphic evidence.

Strobograms may be duplicated for reports by recording on flimsy graph paper with carbon backing paper to emphasize the pen track, and blueprinting the results. The apparent drawback to this arrangement is the loss of various colours that can be used in the original, but there is also a serious conceptual drawback in conceding a residual application for photographic techniques.

Acknowledgments

Thanks are due to the directors of E. K. Cole Ltd for permission to publish this article, and to Messrs. Parkinson and Cowan Limited, who on behalf of La Compagnie pour le production des Compteurs and Matèrials d'usinies à Gaz gave permission to reproduce the illustration of Hospitalier's 'Ondographe', also to those organizations and persons referred to in Part 1 who kindly corresponded on the subject of the history of the method.

Although the description in Part 2 of the article is based on the first prototype of the machine developed by Ekco Electronics Ltd, it should not be interpreted as a specification of that instrument.

* *Courtesy of Nagard Ltd.*

A New Marine Radar

A new marine radar, the 'Escort' type 601 has recently been introduced by the British Thomson-Houston Co. Ltd. This equipment has been designed to make the display interpret the situation instead of merely recording it and to make the handling of the set seem natural and easy. A few of the more important features are described below.

'Chart Plan' Presentation

Four forms of display can be selected to meet different situations. An important one is that in which the centre of the rotating sweep (i.e. the ship's own position) is moved across the picture in the direction of the heading marker, and at a speed corresponding to the ship's own speed. This endows the picture with a fixed frame of reference. Echoes from stationary objects appear as stationary echoes, while echoes from moving objects draw behind them 'tails' which show their aspect and true course. A repeater from a transmitting compass, and a pair of contacts on an electric log (or alternatively an estimate of speed set in by hand) are required to generate the translatory movements which correct for the ship's own movement, and stabilize the frame of reference of the display.

This form of presentation is called 'Chart Plan', because of its similarity in use to plotting on a chart. Its merits arise from the same cause; it gives at a glance information which could be obtained from a conventional 'relative' p.p.i. only by timed observations and the use of tables. The true course of other vessels and the identification of buoys and lightships by their absence of movement may serve as examples.

On the other hand some 'relative' data is valuable; in particular 'closest approach distance', which is readily estimated from a 'relative' p.p.i. by extending the tails of other vessels past 'own ship's position' In order not to lose this facility the second form of Escort display is one in which the Chart Plan shift generators are temporarily arrested, so that the tails of echoes will turn and indicate their relative courses. The other displays are relative; either 'North Upwards' or 'Ship's Head Upwards'.

'Vantage Point' Off-Centring

With Chart Plan presentation the ship's own position moves across the face of the display, and periodically has to be reset. It is started from a point two-thirds of a radius away from the geometric centre of the display, and allowed to move through the centre and approximately one-third of a radius beyond, so that it will require resetting at intervals which depend on the scale of display in use, and the ship's speed.

Three problems arise in connexion with resetting: the right time for it to occur, the choice of position to off-centre to, and the speed of reaching it. In addition to choosing the time and place for resetting, the 'Vantage Point' system performs the operation quickly. The electronic reset reaches the Vantage Point before the aerial has rotated two degrees.

'Auto-Trim' Picture Alignment

Chart Plan presentation requires a compass repeater, and this is left connected when Escort is shut down. It is possible to arrange, therefore, that, on switching on, the heading marker is correctly orientated, regardless of the changes of course since the equipment was last in use. This is done entirely automatically, and the same circuit reorientates the picture 'Ship's Head Up' when that form of display is called for.

Bearing Marker Control

The bearing marker is a radial line, similar to the heading marker, painting once per revolution of the aerial. It is distinguished from the heading marker by the fact that the radius at which it begins to paint is the radius of the indicated range, whereas the heading marker paints from the origin.

This type of marker is notoriously awkward to use, since there is no visible response on the tube face to an adjustment of the control knob, until the aerial sweeps round again and paints in the new position.

The solution found for this problem in Escort is to control the direction of the bearing marker by a mechanical drive taken from the movable graticule in front of the c.r.t. It is arranged to be always parallel to the lines on the graticule, regardless of the point of origin.

Transistor Amplifier for Magnetic Tape and Drum Playback

By A. E. Bachmann*

Magnetic tapes and drums are frequently used as storage or transmission elements, containing the information in a digital form. This information is made available again by playing back the tape or drum through a magnetic head and an amplifier. The article discusses the problems involved in the realization of such an amplifier, when using transistors only. An analysis is given and the design procedure is worked out. A prototype amplifier has been tested, which operates from a playback head having an open-circuit output voltage of more than 2mV. It is capable of driving three bistable multi-vibrators in parallel over a frequency range from 2 to 400 kc/s and a temperature range of 0°C to +70°C. The rise-time of the output pulse is less than 1μsec.

(Voir page 248 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 254)

IN computers a magnetic tape or drum is frequently used as a storage or transmission element. The information, which is most often stored on the tape or drum in a digital form, can be made available again, by playing back the tape or drum through a magnetic head and an amplifier. This article covers an investigation made in order to design such playback amplifiers using generally available transistors. The general requirements are as follows:

In order that the proper head can be chosen from those which are commercially available, the necessary specifications must be calculated. The output of the amplifier must be capable of driving two bistable multivibrators in parallel.

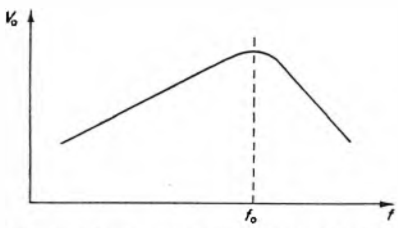


Fig. 1. Magnetic tape playback characteristic

Analysis

Playback Characteristics[1]

The open-circuit output voltage v_o of a magnetic tape playback head is given in Fig. 1.

According to the law of induction the head voltage v_o is:

$$v_o = -(d\Phi/dt) \, . \, w = -w \, . \, \Phi \, . \, f \; \ldots\ldots\ldots \; (1)$$

where w is the number of windings in which the magnetic flux Φ changes according to the magnetization of the tape, here assumed to be of the frequency f. v_o inclines with 6dB/octave up to the frequency f_o. The subsequent fall off is due to self-demagnetization of the tape, the finite width of the gap in the playback head, which cannot be made infinitesimally small, as well as hystereses and eddy current losses in the head.

The equivalent circuit of the head is given in Fig. 2. It consists of an inductance L and a very small loss resistance r. Typical values for such a playback head are: (Potter Model 6404-7, 112 turns) $L = 1$mH, $r = 6\Omega$, $V = 7{\cdot}5$mV. Notice, that the voltage generator V is not constant, but increases with frequency at the rate of 6dB/octave.

If linearity of the playback system is of importance (e.g. in audio—and video playback system) equalizers which linearize the transfer characteristic must be applied.

Signal and Necessary Bandwidth

The information is stored in digital form on the tape, i.e. the tape is either magnetized or not magnetized. Thus when played back, the signal out of the head consists of the presence or absence of a pulse. The tape runs at a speed of 75in/sec for 15msec. It then stops and may be restarted after an unspecified delay. The signal pulses occur at a maximum repetition rate of 15kc/s (90kc/s) or with a minimum spacing of 66·7μsec (11·1μsec)*. The exact waveform is not known but it is assumed that the pulses consist of a sine wave with third harmonic distortion. Because of the digital form, one bit of information consists of a single pulse of an approximately sinusoidal wave. The function $s(t) = \sin(\Omega t + \phi)$ has an amplitude spectrum $|S|$ as given in Fig. 3 for $\phi = 0$, when starting from zero at $t = 0$. In the case of a large third harmonic the spectral distribution will be different but will still have the same general shape. It can be seen that a very broad bandwidth is not necessary

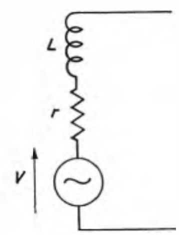


Fig. 2. Playback head

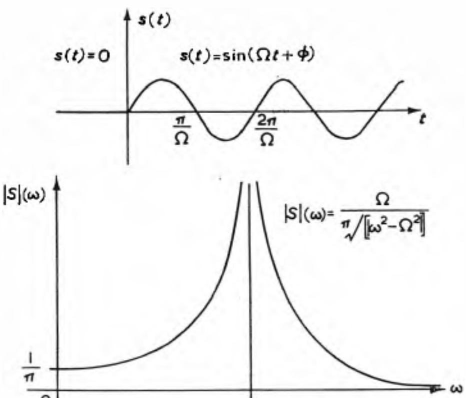


Fig. 3. Amplitude spectrum of a sine wave burst

* PTT Res. Lab., Berne, Switzerland.

* See example under Amplifier Design.

in order to amplify the signal coming from the head. A bandwidth from $\Omega/3$ to 3Ω is thought to be sufficient. Everything beyond that may improve the waveshape, but also decreases the signal-to-noise ratio due to the broader bandwidth.

The Three Amplifier Blocks

The signal coming from the head has quite a large amplitude of noise superimposed upon it. This noise appears because the tape does not run smoothly over the gap of the head, the distance between it and the head changing quite considerably. The signal may be thought of as having the amplitude 1 (see Fig. 4) with a possible fluctuation of $1 \pm a$, and a minimum signal-to-noise ratio of N_{min}, (i.e. the amplitude of the noise is never larger than $1/N_{min}$). Fig. 4 shows that in order to distinguish between noise and signal a threshold circuit must be adjusted to operate in the region

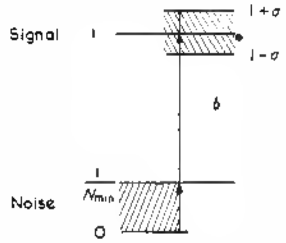


Fig. 4. Signal and noise

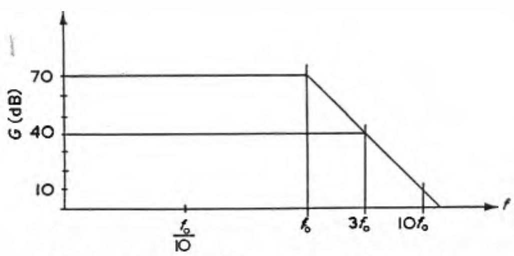


Fig. 5. Gain bandwidth of a three stage amplifier

δ between $1/N_{min}$ and $1 - a$. Since this region δ is very small in the original signal from the source (in the order of millivolts) it is necessary to amplify the signal by an amount A so that an ordinary non-linear element such as a diode can distinguish between the amplified noise $A \cdot 1/N_{min}$ and the signal A. After amplification the signal-to-noise ratio has not been changed (assuming the noise of the amplifier itself is considerably less than the tape noise), but the margin between signal and noise has been increased from δ to $A \cdot \delta$. It can be seen that a system using a fixed threshold level has certain limitations. It only operates as long as the fluctuation $\pm a$ of the signal is not too large. If the variation of signal amplitude is very great, a.v.c. would have to be applied to the bias of the threshold element.

Linear Amplifier and Input Considerations

The linear amplifier is necessary in order to bring the signal up to a level such that it may be easily distinguished from the noise, i.e. The margin $A \cdot \delta$ between signal and noise is made large enough so that a diode threshold element can be adjusted in such a way that only voltages which are higher than a certain specific value are considered as signals, everything lower than that as noise. The linear amplifier must have (a) stable gain and (b) the necessary frequency response.

(a) *Gain Requirements*

Assuming that the output of the linear amplifier is 1V across 3kΩ for an input signal of 7·5mV, the required power gain is:

$$G = (V_o/V_i)^2 \cdot R_i/R_L = 1{\cdot}2 \times 10^4 = 40{\cdot}8\text{dB} \quad \ldots\ldots (2)$$

Output voltage: $V_o = 1$V

Input voltage: $V_i = 7{\cdot}5$mV

Load resistance: $R_L = 3$kΩ

Input resistance: $R_i = 2$kΩ (for common-emitter stage)

Since the gain of a transistor stage is about 25dB it follows that a two stage amplifier will be sufficient. A gain control is necessary for eventual adjustment of the output level. If extremely stable gain is required, feedback would have to be applied.

(b) *Frequency Response*

Applying feedback to the amplifier also helps to increase the bandwidth as can be seen in Fig. 5. A three stage amplifier is shown having a low frequency power gain of 70dB

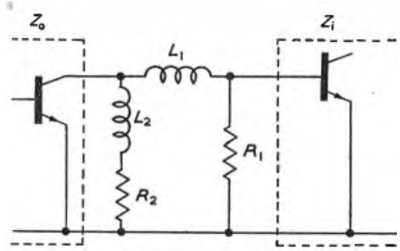


Fig. 6. Series-shunt peaking

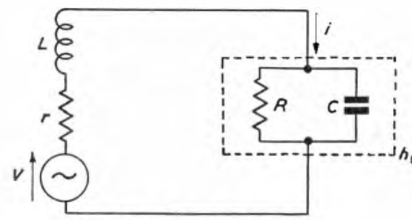


Fig. 7. Input circuit

and the -9dB point at f_o. If the gain of this amplifier is reduced by 30dB with the aid of feedback, then the cut-off frequency is made approximately three times larger. (Notice: a similar change occurs at the low frequency end). To build a three stage feedback amplifier of this type leads to problems of stability. It would have to be designed carefully by taking into account all the phase shifts due to transistors and coupling networks. Using generally available transistors (e.g. 2N123, 2N167) in the common emitter configuration, amplifiers having a flat response up to 100kc/s to 300kc/s can be built. If signals of higher frequencies have to be amplified, the design technique used is that which is adopted for video amplifiers[2]. For convenience Fig. 6 shows an example of interstage networks using series-shunt peaking.

(c) *Input Considerations*

The input circuit, given in Fig. 7, shows the playback head and the transistor represented by its input impedance h_{11e} (R and C in parallel).

The following transfer function $F = (i/V)$ is of interest and is calculated below:

$$F = (i/V) = \frac{1}{r + pL + [R/(1 + pT)]}, \text{ with } p = j\omega \text{ and } T = RC \quad \ldots\ldots\ldots\ldots (3)$$

with the approximation $r \ll R$

$$L \gg CRr$$

equation (3) gives:

$$F = (1/R)\frac{1 + pT}{1 + P(L/R) + p^2CL} = (1/R)\frac{1 + pT}{1 + 2\zeta p\tau + p^2\tau^2} \quad \ldots\ldots (4)$$

from equation (4) it can be shown that

$$\left.\begin{array}{ll} T = RC & f_- = (1/2\pi T) \\ \tau = \sqrt{(LC)} & f_0 = (1/2\pi\tau) \\ 2\zeta = (1/R)\ \sqrt{(L/C)} & Q = (1/2\zeta) \end{array}\right\} \ldots\ldots\ldots (5)$$

the representation of equation (4) is given in Fig. 8.

The peak which occurs at f_0 and which is due to the resonance of L and C, can be used to compensate the losses of the amplifier or head at the higher frequencies. From equation (5) it is possible for a given transistor, having a certain input impedance h_{11e} to determine the maximum permissible head inductance L, so that resonance occurs at the right frequency f_0:

$$L_{max} = \frac{1}{(2\pi f_0)^2 C} \ldots\ldots\ldots\ldots\ldots (6)$$

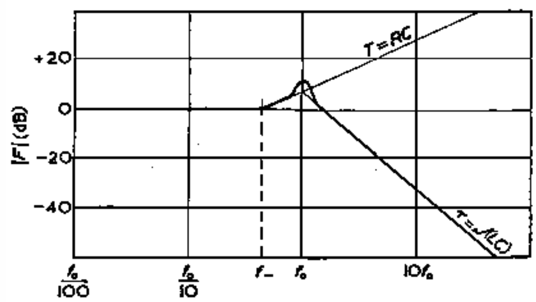


Fig. 8. Transfer characteristic of the input circuit of Fig. 7

The power transfer from the source to the first transistor is a maximum when source and input are conjugate matched, i.e.

$$r - pL = \frac{R}{1 + pCR} \ldots\ldots\ldots\ldots\ldots (7)$$

Since equation (7) must hold for the real as well as for the imaginary part, two equations follow:

$$r = L/CR \ \ldots\ldots\ldots\ldots\ldots\ldots (8)$$

$$r + \omega^2 LCR = R \ \ldots\ldots\ldots\ldots\ldots (9)$$

from equations (8) and (9), the frequency ω_m of conjugate matching can be calculated as:

$$\omega_m^2 = \omega_0^2 (1 - L/CR^2), \text{ with } \omega_0^2 = 1/LC \ \ldots\ldots (10)$$

In a typical case, where $L = 10^{-3}$H, $C = 10^{-9}$F, $R = 2 \times 10^3\Omega$, equation (10) gives:

$$\omega_m = \omega_0 \sqrt{(1 - (1/4))} = 0{\cdot}87 \times \omega_0 \ \ldots\ldots (11)$$

a value which is close to ω_0 itself. From equation (10) a value for R can be calculated:

$$R = \sqrt{\left(\frac{L/C}{1 - (\omega_m/\omega_0)^2}\right)} \ldots\ldots\ldots\ldots\ldots (12)$$

if $(\omega_m/\omega_0) < 1/3$ equation (12) becomes

$$R \simeq \sqrt{(L/C)} \simeq 1\text{k}\Omega \ \ldots\ldots\ldots\ldots (13)$$

So if R can be chosen after equation (13) the frequency ω_m where the source and the input are conjugate matched is appreciably lower than ω_0 where there is resonance between L and C. How much lower it is cannot be said exactly, since control over R is not very good.

Threshold Element

The purpose of the threshold element is to pass only signals of greater than a certain magnitude, say C and block everything which is of a lower level. Fig. 9 shows a sinusoidal signal $V_A = V_{AO} + 2\,C\sin(\omega t + \phi)$ as it appears at the collector of an npn transistor, Fig. 10 showing the circuit. As long as the potential V_A of the collector is less than the d.c. potential V_{BO} of the voltage divider R_1, R_2, the diode D is biased backwards and no signal current flows into the next stage. The amplifier is blocked. As soon as the potential V_A of the collector exceeds V_{BO}, signal current

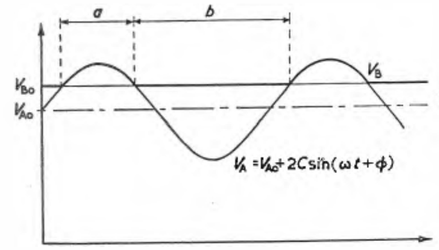


Fig. 9. Thresholding a sine wave

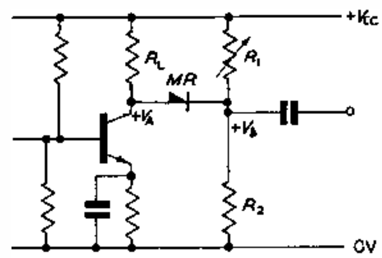


Fig. 10. Threshold element

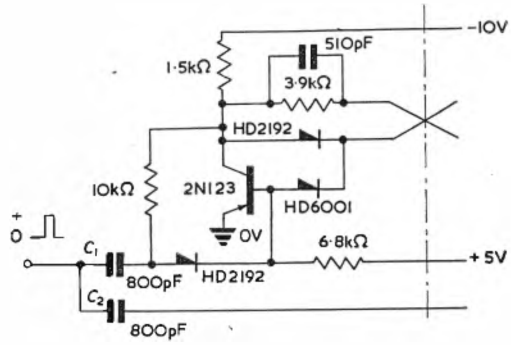


Fig. 11. Half of bistable multivibrator

flows into the next stage. The amplifier is unblocked. The amount of threshold C can be adjusted by varying the d.c. potential V_{BO} with the aid of the potentiometer R_1. As long as the magnitude of the signal is large enough, say larger than 0·5V, the characteristics of the diode are not important. If a smaller signal has to be thresholded (e.g. because the signal from the source is appreciably lower than expected) a high back resistance diode (e.g. HD2085) will give sharper clipping.

Non-linear Amplifier and Output Consideration

At the output, two bistable multivibrators of the type shown in Fig. 11 have to be triggered in parallel by means of positive triggers. To do so, the output must be capable of delivering a pulse of approximately 10V and 10mA with a rise-time of 2 to 3μsec. Since the length of the pulse is not very critical (it will be differentiated by the coupling capacitors C_1 and C_2 of the multivibrators) the non-linear amplifier can be designed with non-clamped saturated stages. The rise-time t_r of the common-emitter pulse amplifier in Fig. 12 is given by[2]:

$$t_r = \frac{1}{(1 - \alpha_{fb})\cdot\omega_{\alpha fb}} \cdot \ln\left[\frac{1}{1 - \dfrac{0{\cdot}9\,(1 - \alpha_{fb})}{\alpha_{fb}}(I_c/I_b)}\right] \quad (14)$$

where α_{fb} = forward α_b

$\omega\alpha_{fb}$ = α_b cut-off frequency in forward direction

I_b = driving current at the base

I_c = saturation current

Thus for a given transistor, the rise-time can be made shorter by decreasing the saturation current I_c and increasing the driving current I_b at the base.

Amplifier Design

After the analysis of the amplifier, the design of any particular playback amplifier will be relatively simple. For the purpose of this development, the following two types are

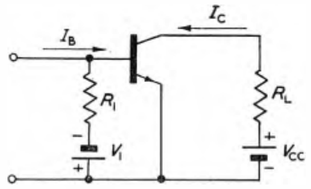


Fig. 12. Common emitter pulse amplifier

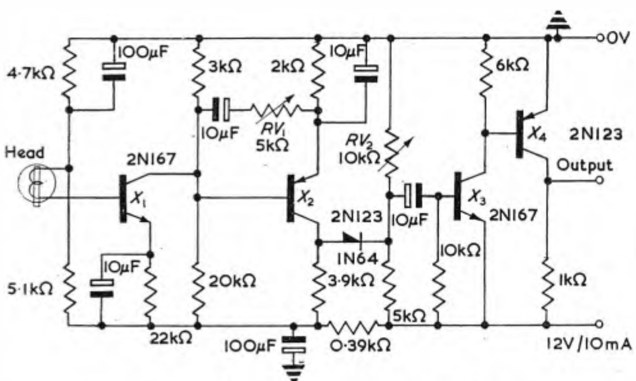


Fig. 13. Magnetic tape playback amplifier

of interest:

Type	(a) Tape Playback	(b) Drum Playback
Pulse repetition rate	15kc/s	90kc/s
Head inductance	1mH	50μH
Head loss resistance	6Ω	
Open-circuit output voltage	7·5mV r.m.s.	20mV
Number of 'flip-flops' to be driven in parallel	2	
Threshold level	50 per cent	

First equation (6) has to be checked in order to find out if the first stage can be an ordinary common-emitter stage for the given heads. Such a stage has a shunt capacitance C of approximately 0·001μF. If the resonance frequency is chosen as three times the pulse frequency, it can be seen that in both cases the given inductance of the head is less than the required L_{max} given by equation (6).

Both amplifiers can be built without using feedback or peaking networks. As outlined in the analysis two stages are sufficient for the linear amplifier. This will give some reserve of gain in case (a) and more in case (b). Fig. 13 shows the circuit which can be used for both cases. This circuit has actually been built and tested.

The linear amplifier makes use of the npn, pnp symmetry for direct coupling. Potentiometer RV_1 is the volume control. Potentiometer RV_2 sets the threshold value. To set the threshold level the following procedure is recommended:

(1) Feed in the amount of signal available as a sine wave.

(2) Adjust the potentiometer RV_1 so that the signal level at the collector of transistor X_2 is in the order of 1V r.m.s., (i.e. for input signals below 20mV r.m.s. put RV_1 at maximum).

(3) Look at the waveform after the diode on an oscilloscope and adjust the threshold potentiometer RV_2 so that proper thresholding is achieved. The ratio of b/a in Fig. 9 tells how much of the sine wave comes through (e.g. 50 per cent for $b/a = 2$).

Assuming that the input current into the base of X_3 is approximately 0·25mA and the short-circuit current gain α_e

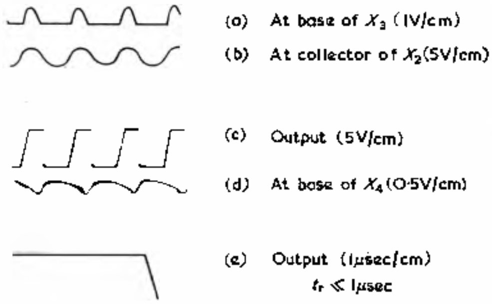


Fig. 14. Waveforms

Input: 7·5mV r.m.s. sine wave of 15kc/s
Volume (RV_1) at maximum
Threshold (RV_2) at 50 per cent

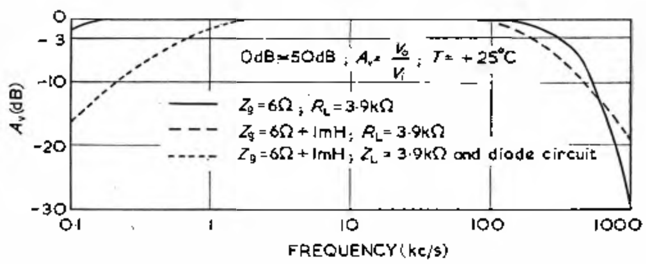


Fig. 15. Frequency response of the linear amplifier

of each transistor is 20, the collector current of X_3 is 5mA. Since the saturation current of X_3 is 12V/6kΩ = 2mA, 3mA flow into the base of X_4, giving a collector current in X_4 of 60mA. Deducting the saturation current of 12mA, this means that approximately 48mA are available to drive the flip-flops.

For applications where repetition rates up to 100kc/s may be encountered the last two stages can also be made direct coupled. Since the storage time is too long for higher repetition rates, a simple differentiating RC network must be inserted. With this addition the amplifier operates with repetition rates up to 400kc/s.

Measurements

Fig. 13 gives the final circuit of the amplifier as it has been designed and tested.

Fig. 14 shows the different waveforms at certain points of the amplifier when the input signal is a sine wave of 15kc/s.

In Fig. 15 the frequency response of the linear amplifier is given. The playback head was represented by a coil of 1mH inductance. Resonance between this inductance and the input capacitance of the first stage occurs at about 300kc/s.

Fig. 16 gives the rise-time as a function of the amplitude of the input signal. A relative amplitude of 1 means that amount of input signal (e.g. 7·5mV r.m.s.) such that 50 per cent is cut off by the threshold element. In an ideal case any signal having a magnitude of less than 0·5 should not come through, thus t_r would be infinite. Since the diode characteristic is not ideal and the average of the impedance looking into the diode circuit depends on the threshold level, infinite rise-time occurs at a relative amplitude of 0·3 rather than 0·5.

In Fig. 17 the voltage gain of the first two stages is plotted as a function of temperature. It is flat within 3dB from +10°C to +70°C. Below 0°C it drops off more rapidly.

The amplifier was capable of driving in parallel, three bi-stable multivibrators of the type shown in Fig. 11. It worked over a temperature range from 0°C to +70°C. Using a sine wave input, it was possible to trigger the multi-vibrators from 2kc/s to 400kc/s. Interchanging transistors did not affect the proper working of the amplifier.

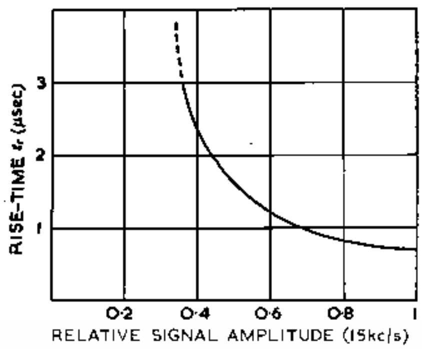


Fig. 16. Rise-time against input amplitude

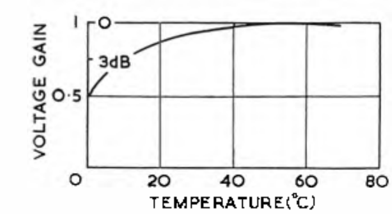


Fig. 17. Linear voltage gain against temperature

Acknowledgments

The author is indebted to the Electronics Laboratory of the General Electric Company, Syracuse, N.Y., where he had the opportunity of carrying out the work described.

Appendix

Equalized Linear Amplifier

The foregoing investigation is based on the assumption that all the pulses appearing at the input of the amplifier (base of X_1) have the same amplitude within the tolerated fluctuation limits $\pm a$. This is necessary if the information to be transmitted consists of both single pulses and sequences of pulses. When a fixed threshold element is used, it is very important that each pulse reaches the diode with the same amplitude so that the threshold level for all of them will be the same.

If the above assumption does not hold and the amplitude of the pulses from the playback head are dependent on the repetition rate, the response of the linear amplifier has to be shaped by means of equalizing networks.

First the playback characteristic of the system involved (drum or tape) has to be measured in order to find the slopes of the response and the frequency f_o as indicated in Fig. 1. Once this is known, an equalizing network can be found which together with the amplifier gives a flat overall response.

A typical example is that shown in Fig. 18, where the transistor is by-passed with an *RC*-network. The transfer function $F = i_2/i_1$ given by

$$F = i_2/i_1 = \frac{Z}{Z + h_{11e}} = \frac{(1 + pT_1).(1 + pT_2)}{(1 + pT_1).(1 + pT_2) + pR_2C_1} \quad (15)$$

where $T_1 = C_1R_1$; $T_2 = C_2R_2$; $p = j\omega$. In order to make F the image of the playback characteristic over the necessary frequency range from $f-$ to $f+$ it can be shown that the following equations must hold:

$$R_1C_1 = R_2C_2 = T_o = 1/2\pi f_o \quad \ldots\ldots\ldots \quad (16)$$

$$C_1 \simeq KC_2, \text{ where } K = f_o/f- \quad \ldots\ldots\ldots \quad (17)$$

R_2 is given by the transistor and its bias resistors. From equation (16) C_2 can be found. Part of C_2 will consist of the input capacitance of the transistor. The choice of K depends on the system to be used and can be as high as 100. Equation (17) gives the value

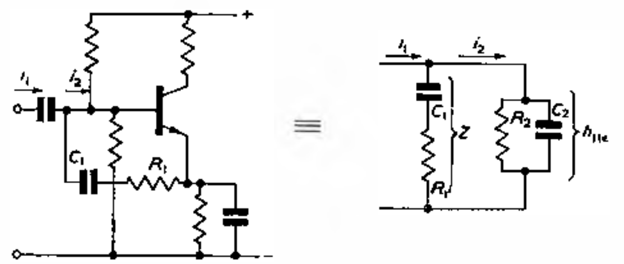


Fig. 18. Equalizing stage

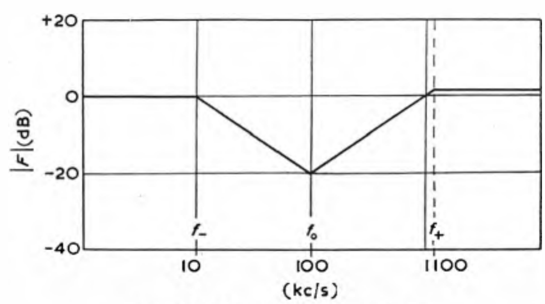


Fig. 19. Equalizing characteristic

of C_1 from K and C_2 and equation (16) again gives R_1 from C_1 and T_o.

Example: $R_2 = 2\text{k}\Omega$

$f_o = 100\text{kc/s}$

$f- = 10\text{kc/s}$

The calculations lead to:

$C_2 = 800\text{pF}$

$K = 10$

$C_1 = 0{\cdot}008\mu\text{F}$

$R_1 = 200\Omega$

Fig. 19 shows the absolute value of the transfer function $F = i_2/i_1$ for the above example. The upper frequency $f+$ is given by $f+ = 1/2\pi\tau_2$, where τ_2 is

$$\tau_2 = T_o + (R_2C_1/2) - \sqrt{[T_o R_2C_1 + (R_2C_1/2)^2]} \simeq 0{\cdot}14\mu\text{sec} \quad \ldots\ldots\ldots\ldots \quad (18)$$

$$f+ = 1/2\pi\tau_2 \simeq 1{\cdot}1\text{Mc/s} \quad \ldots\ldots\ldots \quad (19)$$

Since the equalizing network introduces a loss, additional stages have to be used in the amplifier in order to meet the gain requirements.

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A Simple Very Low Frequency Oscillator

By J. F. Young*, A.M.I.E.E., A.M.Brit.I.R.E.

The usual methods of amplitude stabilization of resistance-capacitance oscillators are frequency sensitive. In the simple oscillator described a Zener diode limits the amplitude and the resulting harmonics are filtered out by a selective circuit which also controls the frequency of oscillation. A reasonable waveform can therefore be obtained even at very low frequencies.

(Voir page 248 pour le résumé en français; Zusammenfassung in deutscher Sprache auf Seite 254)

It is usual to add some form of amplitude limitation to valve oscillators in which the frequency is controlled by resistance-capacitance selective circuits. This is required for two reasons, firstly in order to maintain a constant output voltage irrespective of load variations, and secondly to prevent build-up of oscillation to the point where overloading causes excessive harmonic distortion to appear. These are often conflicting requirements, since when the load is heavy, a system designed to maintain a constant mean or r.m.s. value of output voltage can cause amplitude distortion to appear. The two requirements should therefore be met by separate control circuits. In many cases it is not necessary to have a low output impedance and amplitude limitation is only required in order to avoid distortion.

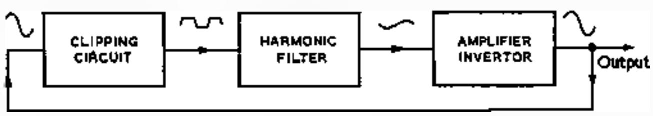

Fig. 1. Block diagram of oscillator

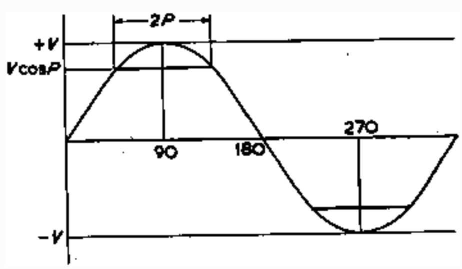

Fig. 2. Clipped sine wave

The methods generally used fall into two classes. The first class uses some form of feedback automatic gain control, in which the voltage at some point in the oscillator is measured and used to control the gain of the oscillator loop. The a.g.c. action is of necessity slow, and difficulties are encountered where the oscillator must cover a wide frequency range.

The second class of amplitude limitation makes use of thermal non-linear elements, such as lamps or thermistors, in which the slope resistance depends on the r.m.s. value of the current flowing through the device. These make very simple circuits possible, but they have two disadvantages. The fact that they are thermal devices makes them sensitive to ambient temperature, though this can be allowed for in the design. A more serious disadvantage is that unless special steps are taken to slow down the response, thermal elements are not suitable for use in v.l.f. oscillators since they would respond to instantaneous rather than r.m.s. amplitudes.

Since existing methods of stabilization are frequency sensitive, the possibility of using an amplitude limiter (or clipper) is worth exploring. Such a clipper is not sensitive to frequency, but it produces harmonic distortion. However, the selective circuit of an oscillator can be used both to control the oscillation frequency and to filter out harmonics produced by the clipping circuit. A block diagram of an oscillator utilizing this principle is shown in Fig. 1. Here the output sine wave is fed back and passed through the clipping circuit where the peaks are removed. The filter then removes most of the harmonic content and the resulting sine wave is amplified.

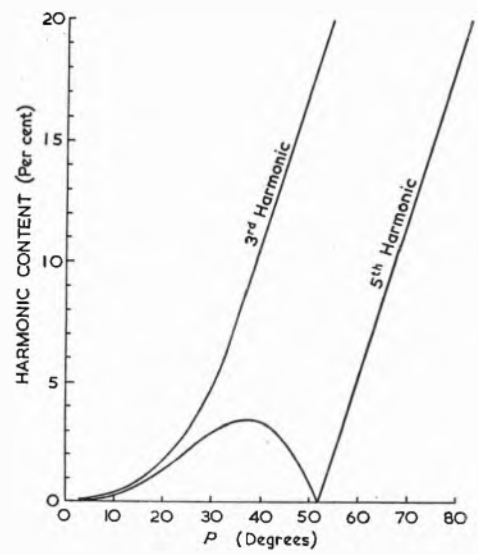

Fig. 3. Harmonic content of clipped sine wave

A sine wave $V \sin L$ is shown in Fig. 2 clipped at an amplitude of $V \cos P$, the angle of clipping being $90° \pm P$.

Fourier expansion of this waveform gives:—

$$\begin{aligned} 2V/3{\cdot}14\ [&(3{\cdot}14/2 - P + \tfrac{1}{2}\sin 2P)\sin L \\ &+(\tfrac{1}{2}\sin 2P - \tfrac{1}{4}\sin 4P)\,\tfrac{1}{3}\sin 3L \\ &-(\tfrac{1}{4}\sin 4P - 1/6\sin 6P)\,1/5\sin 5L \\ &+ \ldots\ldots\ldots\ldots] \quad \ldots\ (1) \end{aligned}$$

The percentage third and fifth harmonics are plotted in Fig. 3. It will be noticed that the 5th harmonic disappears

* *The General Electric Co. Ltd, Birmingham.*

completely at approximately $P = 52°$ when the wave is clipped down to about 0·6 of its peak amplitude. The third harmonic is then about 16 per cent of the fundamental, or 16dB down.

As the amount of clipping increases, the coefficients approach those for a square wave of very small amplitude, as would be expected.

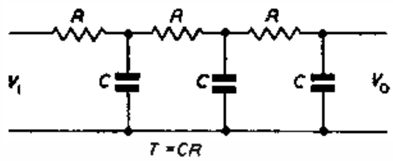

Fig. 4. Selective circuit

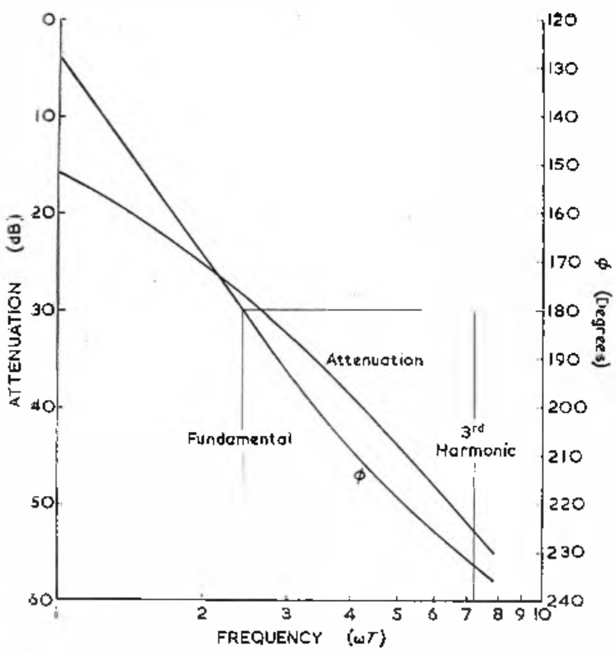

Fig. 5. Response of selective circuit

In order to obtain a reasonable waveform, the third harmonic content of the clipped wave must be attenuated by some tens of decibels below the fundamental. This cannot be done with the Wien bridge or the parallel-T networks commonly used in oscillator circuits since these give a ratio of only about 6dB between the frequency of oscillation and the third harmonic. Multiple ladder networks can give better results, and can be used in quite simple oscillator circuits.

The ladder circuit shown in Fig. 4 has a frequency response of the form:—

$$V_o/V = \frac{1}{1 - 5\omega^2 T^2 + j\omega T(6 - \omega^2 T^2)} \quad \ldots\ldots \ (2)$$

It can be used in an oscillator circuit, since it gives a 180° phase shift at a frequency where $\omega T = 2{\cdot}45$. The response is plotted in Fig. 5. The attenuation at $\omega T = 3 \times 2{\cdot}45$ is about 24dB greater than that at the fundamental frequency, and the resulting third harmonic distortion when used in conjunction with the clipped wave mentioned would be 40dB down, or 1 per cent.

The type of selective circuit mentioned has been widely investigated and used in oscillators, and there are many references to it[1]. Some preliminary tests on such a circuit were carried out with the arrangement of Fig. 6 working at 50c/s. An EW77 Zener diode was used to clip the waveform. This has the advantage that it can provide an effective 3V bias for the grid of a valve at the same time as it clips the waveform, so there is no necessity for other biasing arrangements.

The output amplitude was reasonably constant at about 0·23V peak-to-peak for input amplitudes down to about 10V peak-to-peak

The amplifier gain required for use in an oscillator is a minimum of about 30dB, as will be seen from Fig. 5. The amplifier must be designed so that it will not saturate with

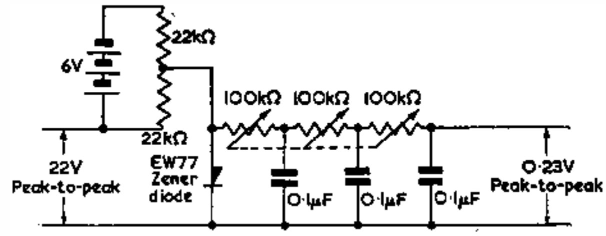

Fig. 6. Test circuit

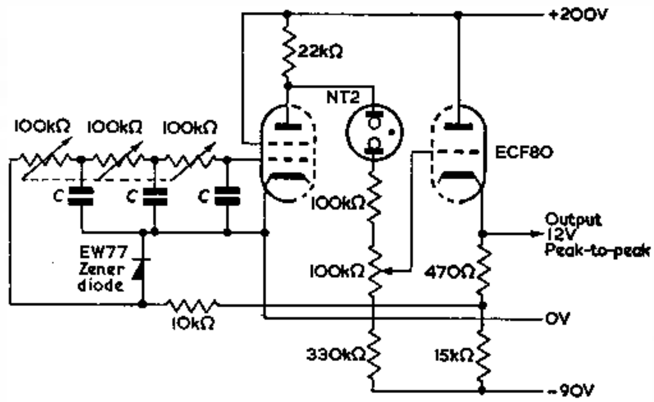

Fig. 7. The simple oscillator

several times 0·23V peak-to-peak input, while providing more than the specified 10V peak-to-peak output. This is not difficult, and a suitable circuit is given in Fig. 7.

The circuit uses an ECF80 triode-pentode valve as an amplifier and cathode-follower. It is entirely d.c. coupled so that there is heavy d.c. negative feedback in the absence of the Zener diode. This ensures that the Zener diode cannot be biased to a condition where it can prevent the build-up of oscillation. The Zener diode provides the bias for the pentode section and clips the waveform. The output obtained is about 12V peak-to-peak, and since it is taken from the cathode-follower, the output impedance is reasonably low. No difficulty is encountered in starting oscillation and there are no transient effects when tuning.

More complicated versions of this basic arrangement could give improved results. These would involve the use of a four gang tuning resistor, or preferably capacitor, and an extra cathode-follower to reduce the source impedance seen from the selective circuit. The latter addition would make it possible to reduce the loading of the output cathode-follower by the clipping circuit. In the simple circuit it is necessary to compromise to keep the clipping circuit source impedance low, while not excessively loading the cathode-follower by the Zener diode.

If the variable resistors are made too small compared with the source impedance, oscillation ceases, since the gain is reduced. Variation of the capacitors rather than the resistors, removes this effect, as does the addition of fixed resistors in series with the variable ones to provide a definite minimum value.

The circuit is based on the deliberate introduction of amplitude distortion at a point in the loop where it can have only a small effect on the output waveform. In the normal oscillator of this type, the amplitude distortion which occurs, accidentally, if no amplitude stabilization is employed, is in such a position that the output is distorted. The circuit given is of particular use as a very low frequency generator. More complicated versions should have a wider use, having the advantage over other methods of amplitude stabilization that transient effects are avoided. It should be noted that there is no attempt to maintain the output voltage constant, and this will vary if the gain of the valves varies. However, this will have little effect on the output waveform.

The circuit shown oscillates at frequencies from 100kc/s down to one cycle per minute. The upper limit is caused by stray capacitance, while the lower limit is mainly determined by the physical size of the capacitors required. If still lower frequencies are required, the common ends of the capacitors in the selective circuit should preferably be taken to a slightly negative voltage. This would avoid the heavy current which would otherwise be taken by the pentode while the capacitors were initially charging up to the correct negative mean running voltage. It would also make the oscillation build up more rapidly after switching on.

Acknowledgment

The author wishes to thank the General Electric Co. Ltd for permission to publish this article.

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1. Shaw, J. C. The Design of Phase Shift Oscillators and Filters (Bibliography). *Electronic Engng. 28*, 216 (1956).

A Stroboscopic Method of Making Frequency Response Measurements on Small Electromechanical Devices

By M. Shepherdson*, B.Sc. and R. Walters*, A.M.I.Mech.E.

A method is described for making frequency response measurements on small electromechanical devices using a travelling microscope and a stroboscopic lamp. Apart from a simple pulse amplifier and phase shifting network, which is described, the apparatus needed is available in most well equipped electrical laboratories. The method involves no loading of the test component and allows amplitude measurements well below ·001in to be taken.

(Voir page 248 pour le résumé en français; Zusammenfassung in deutscher Sprache auf Seite 254)

FREQUENCY response measurements on small electromechanical components are complicated by the fact that it is impossible to attach a pick-off to the moving member without loading it mechanically. Inductive or capacitive proximity meters can be used which exert a negligible force on the moving part, but these require calibration on each test piece and usually involve the complications of a modulation system. An optical method of measurement has been used which has the advantages of simplicity, accuracy and flexibility. Apart from a simple pulse amplifier and phase shifting network the apparatus needed is available in most well equipped electrical laboratories. In a typical application the responses of all the moving parts of an electro-hydraulic servo system were determined. These included a torque motor driving a pilot valve, the main valve and a hydraulic jack. Amplitudes at the pilot valve were measured down to 0·001in and the jack movements were up to 0·5in. Frequencies were from 10c/s to 250c/s. It is only necessary that the part undergoing test be visible.

The Method

The test component is excited at the desired frequency from an oscillator through an appropriate amplifier and the moving element examined through a travelling microscope with cross hairs or an eyepiece scale. Light is provided by a stroboscopic lamp triggered by the oscillator. Between the oscillator and lamp a phase shifter is interposed, capable of continuous phase shift up to 360°, also a pulse forming amplifier. In the latter the sinusoidal input is squared and differentiated to give a sharp pulse each time the signal passes through zero. The circuit is arranged to apply all the pulses to the stroboscope in the same sense, thereby giving two flashes per cycle, but a switch permits the suppression of every second pulse.

Using one flash per cycle, the operator can see one stationary image in the field of the microscope. If he turns the phase control steadily through 360° the image goes through one cycle of movement, passing through the centre twice, in opposite directions, and reaching each extreme once. Under double flash illumination two images are seen. In this case when the phase control is advanced through 360° the two images merge into one twice and reach extreme separation twice. The amplitude can be measured directly, being half the maximum separation of the images. For small amplitudes, the eyepiece scale is used together with a high power objective if necessary. Measurements of large amplitudes are made with the calibrated traverse mechanism of the microscope.

The phase of the motion relative to the oscillator is read off the calibrated phase shifter when the images are in coincidence. There is an ambiguity of 180° owing to the fact that the images merge twice per rotation of the phase

* *Sperry Gyroscope Company Limited.*

shifter. If the performance of the device is not already known with sufficient accuracy to resolve the ambiguity, the correct coincidence must be selected after examination under single flash illumination. The two nulls can be distinguished under single flash operation, because at one null as the phase is advanced the image moves in one direction whereas at the other null advancing the phase moves the image the other way. Reference to the very low frequency performance will reveal which is the correct null to use.

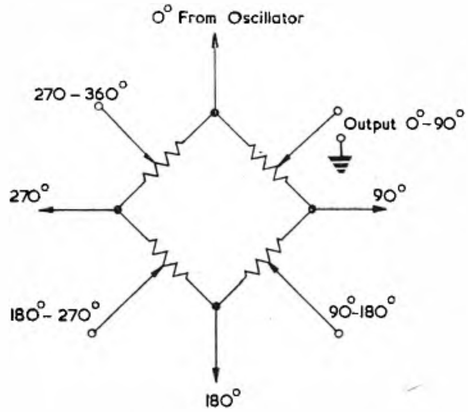


Fig. 1. Phase shifter

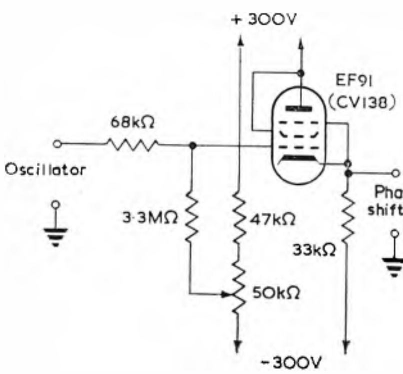
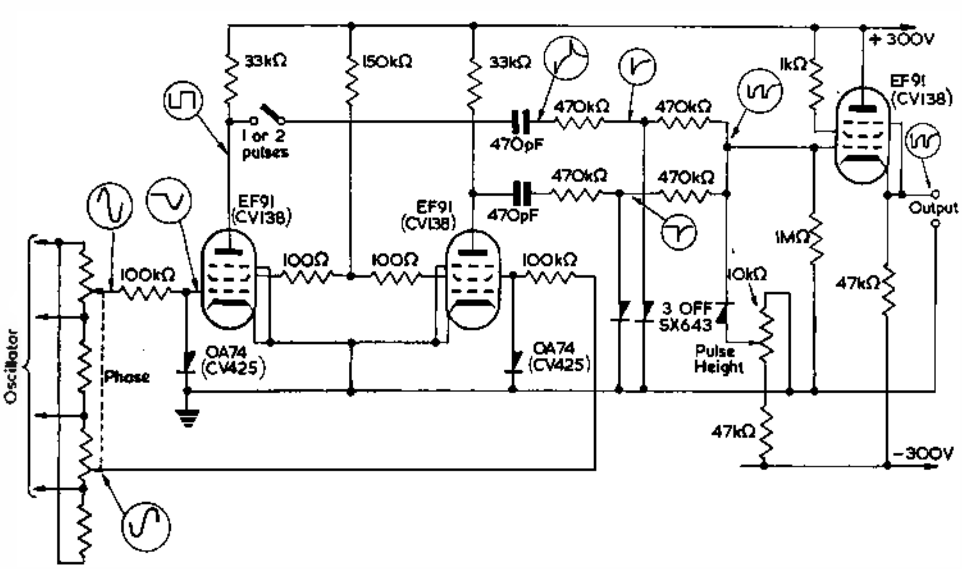


Fig. 2. Amplifier circuit

Fig. 3. Isolating amplifier



Electronic Circuits

In order to make a phase shifter independent of frequency it is necessary to have a polyphase source. The Solartron transfer function analyser has a four phase oscillator which has been found to be suitable. Four potentiometers are connected between the line terminals as shown in Fig. 1. Each potentiometer has a range of 90° and a four way switch selects the desired quadrant. It can be shown that such a phase shifter obeys the law

$$X = \frac{1}{1 \mp \cot \phi}$$

where X is the fraction of the potentiometer taken out and ϕ is the phase shift.

The Solartron oscillator has two terminals at a high impedance level. To prevent loading the oscillator and to retain balance, four isolating stages as shown in Fig. 2 were interposed between the oscillator and the phase shifter.

Fig. 3 shows the amplifier circuit used. A push-pull signal is taken from the phase shifter using a ganged pair of potentiometers. This is applied to the grids of a pair of valves run to saturation, positive grid current being diverted through diodes. The square waves are differentiated and the positive pulses clipped off before addition at the grid of the output valve. A further limiter is included to cut all the pulses to a uniform height of about 50V. These pulses were used with a Dawe Stroboflash.

Conclusion

In order to achieve accuracy with the method described the pulses must have a fast rise-time and be accurately placed on the timing wave. If they are not exactly 180° apart it becomes impossible to bring the two images to coincidence and there is a small error in amplitude measurement. Any delay in the lamp triggering circuit introduces errors to the extent that it is indeterminate. A photocell and amplifier capable of displaying the light flashes on an oscilloscope have been found to be very useful for measuring time lags in the system. With the apparatus described a lag of about 0·5msec was detected and allowed for in the higher frequency measurements.

The lower limit of amplitude that can be detected is governed by microscope technique. Down to about ·001in minor surface blemishes on the moving part provide features on which measurements can be made. Below this level special fine marks must be provided possibly by polishing and etching. When a high power objective is used to measure very small amplitudes the system would be improved by introducing the light through a vertical illuminator. The upper frequency limit is determined by the duration of the lamp flash. A high energy short duration flash not only allows a fast repetition rate by also gives sharp focusing.

REGULATED POWER SUPPLIES

By D. J. Collins*, B.Sc.(Eng.), A.M.I.E.E., and J. E. Smith*, B.Sc.

Regulated, or stabilized, d.c. power supplies are required for a number of purposes. In this article the various types of regulators are briefly reviewed and the closed loop series stabilizer is dealt with in some detail. The design problems of this type of unit are considered and particular reference is made to the loop stability. In an Appendix the design of a typical unit is detailed.

(Voir page 248 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 254)

AS the use of precision circuits increases, the necessity for better power supplies becomes apparent. At some time, the need may be felt for a d.c., h.t., supply having one or more of the following features:

(a) Good regulation, i.e. low internal resistance and, in some instances, low internal impedance.
(b) Good stability, i.e. stable output voltage despite mains fluctuations, temperature changes, etc.
(c) Low ripple and noise content.
(d) Ability to vary the h.t. voltage without detriment to any of the above features.

The demand has been met with a variety of devices, some of which are termed regulated power supplies and others stabilized power supplies. Many of the devices provide both good regulation and good stability, and throughout this article the term regulated supply will be used to imply good regulation or good stability.

Definitions

At this stage, it is necessary to define some of the terms used in specifying a regulated power supply.

OUTPUT VOLTAGE

Although this seems rather obvious, the designer will require to know if variable supplies are required, and if not, to what accuracy the supply voltage is required to be set. It is also necessary to know if the supply will be 'floating', or will have a terminal connected to ground.

OUTPUT CURRENT

The current range must be specified, and the supply must cater for the peak current conditions and not merely for the quiescent level.

SOURCE RESISTANCE

This is the internal resistance of the supply which, in some designs, can be made as low as ·001Ω.

SOURCE IMPEDANCE

This is the internal impedance of the supply, and it is usual to stipulate some frequency limit.

RIPPLE AND NOISE LEVEL

This is usually specified in millivolts peak-to-peak. Quite often, the content has a complex waveform, and an oscilloscope measurement is to be preferred.

STABILITY

Many factors will cause output voltage variations. The most obvious is possible variations in the mains voltage, and it is usual to specify a stabilization ratio or factor which is defined as:

$$\frac{\text{Percentage change in mains voltage}}{\text{Resultant percentage change in output voltage}}$$

POWER INPUT

The overall range of mains voltage and frequency must be specified, and it is usual to design for a limited voltage variation on a nominal input. This limited variation is usually taken as ±7 per cent on electronic supplies, this figure being based upon valve manufacturers' permissible heater voltage variation.

MISCELLANEOUS

Other parameters or features which are required to be specified include environmental conditions, metering, fusing, switching, and connexions.

Simple Stabilizing Elements

The first attempts to provide regulated supplies were made using gas-filled stabilizer tubes. The tubes provide a

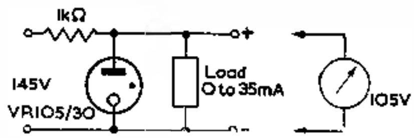


Fig. 1. Simple gas filled regulator

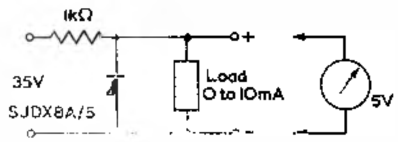


Fig. 2. Simple Zener regulator
Source resistance 20Ω
Stabilization factor 7 : 1
Ripple reduction 50 : 1

simple solution, but have limitations. In general, only supplies having voltages greater than about 70V and currents less than 50mA can be handled in this manner.

A typical circuit is shown in Fig. 1. This circuit provides a source resistance of approximately 60Ω, has a stabilization factor of 14:1, and provides a ripple reduction of 20:1.

Of recent years, a lower voltage element, the Zener diode[1] has achieved some prominence. These diodes make use of the breakdown state of a silicon junction diode when subjected to a reverse voltage. The slope resistance in the breakdown condition is dependent upon the turnover voltage of the particular diode. The temperature coefficient, which is significant, varies from a negative value to a positive value, the crossover point occurring in the 5V region. Improvement in temperature coefficient can be effected by using a combination of diodes but with resultant increase in slope resistance.

Fig. 2 shows the circuit of a Zener regulator, the parameters of which are detailed. This type of circuit is finding favour as a reference level for use in transistor power supplies.

There are available a number of non-linear components, e.g. thermistors, which can be used as stabilizing elements in power supplies.

* *Solartron Research and Development Ltd.*

Hard Valve Voltage Regulators

Hard valve circuits have overcome many of the disadvantages of the gas-filled stabilizer. In general, the control is carried out by the hard valve circuit utilizing the stabilizer tube purely as a stable voltage reference.

There are two main types of circuit configuration, the shunt type and the series type, the latter being more common. The series type gives a lower source resistance, and is more suitable for high and variable voltage supplies. The shunt regulator is more suitable for constant voltage, fixed load supplies. Fig. 3 shows a simple shunt regulator, and indicates the basic parameters.

In view of its wider application and superior performance, only the series type circuit will be considered in detail.

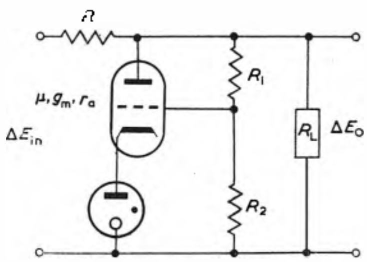


Fig. 3. Simple shunt regulator

$$R_o = \frac{R}{1 + (R/r_a) + \beta R g_m}$$

$$\frac{\Delta E_{in}}{\Delta E_o} = 1 + (R/R_L) + (R/r_a) + \beta g_m R$$

$$\beta = \frac{R_2}{R_1 + R_2}$$

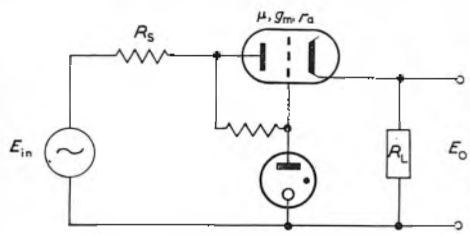


Fig. 4. Simple series regulator

Simple Series Regulator

The series regulator is basically a simple cathode-follower, to the input of which a stable reference level is applied (see Fig. 4). Assuming a perfect reference source, i.e. constant voltage and zero internal resistance, this configuration gives a source resistance R_o, which is the source resistance of a cathode-follower,

$$R_o \simeq \frac{r_a + R_s}{\mu}$$

The stability figure $\Delta E_o/\Delta E_{in} = \dfrac{R_s + r_a}{R_L} + (1 + \mu) \simeq \mu$

when μ is large.

The circuit has the limitation of having its output closely tied to that of the reference. From the formulae, it is apparent that valves having high μ and low r_a give better performance. A typical valve, the A2134, has $\mu = 10$ and $r_a = 835\Omega$. Valves designed for this application are required to operate at high currents with low anode voltages, and in more complex regulators this is of prime importance.

For a supply having a source resistance of less than 10Ω, the single valve cathode-follower is not adequate. Parallel combinations can produce lower figures, but there are more economic ways of producing a better performance.

Closed-Loop Regulator

The performance of the simple cathode-follower can be vastly improved by means of closed loop feedback with a d.c. amplifier. A simple closed-loop system is illustrated in Fig. 5, together with its equivalent circuit.

In analysing the performance of this circuit, the following assumptions are made:

(a) That the gain of the pentode amplifier is $-M$.

(b) That the grid-cathode signal appearing at the series valve is V_g.

(c) That the current drain in the divider chain, R_1, R_2 is negligible compared with the load current.

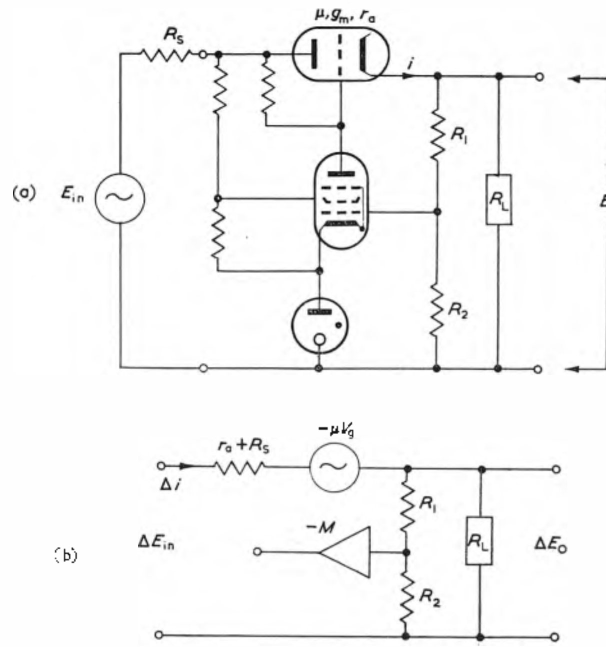


Fig. 5. Improved series regulator

(d) That the transmission of the chain is

$$\beta = \frac{R_2}{R_1 + R_2}$$

Consideration of the equivalent circuit gives the following equations:

$$\Delta E_{in} = \Delta i\,(r_a + R_s) - \mu V_g + \Delta E_o \quad \ldots\ldots\ldots \quad (1)$$

$$V_g = -M\,(\beta \Delta E_o) - \Delta E_o \quad \ldots\ldots\ldots\ldots \quad (2)$$

$$\therefore \Delta E_{in} = \Delta i\,(r_a + R_s) + \mu \Delta E_o (1 + \beta M) + \Delta E_o \quad \ldots \quad (3)$$

From equation (3):

$$\Delta i = \frac{\Delta E_{in} - \Delta E_o\,[1 + \mu\,(1 + \beta M)]}{r_a + R_s} \quad \ldots\ldots\ldots \quad (4)$$

But $\Delta E_o = \Delta i R_L$

Substituting in equation (4)

$$\Delta E_o = \frac{\{\Delta E_{in} - \Delta E_o\,[1 + \mu\,(1 + \beta M)]\}\,R_L}{r_a + R_s}$$

$$\therefore \Delta E_{in}/\Delta E_o = \frac{r_a + R_s}{R_L} + \Big(1 + \mu\,(1 + \beta M)\Big)$$

In practice, the stability figure $\Delta E_{in}/\Delta E_o$ approximates to $\beta M \mu$

Thus, the introduction of an amplifying stage has improved the stability figure by a factor, βM. βM is usually large and is termed the 'loop gain' of the system.

It can also be shown, that for this type of system

$$R_o = \frac{r_a + R_s}{1 + \mu(1 + \beta M)} \simeq \frac{r_a + R_s}{\beta M \mu}$$

and thus the source resistance of the series valve has also been decreased by the factor βM. It is obvious that R_s is not usually purely resistive, since the power source can contain a choke and/or capacitor filter.

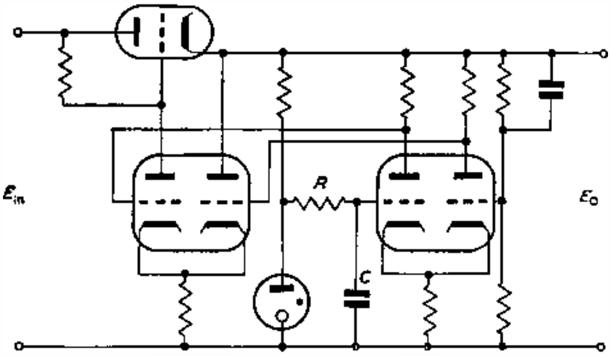


Fig. 6. Series regulator with improved loop gain and a degree of drift correction

type shown in Fig. 5, having an output voltage 250V, dependent upon an 85V reference tube, then this signal produces an output variation of approximately 0·3V or 0·1 per cent of the output. This then implies a maximum stabilization factor of 100 : 1.

Some of the above disadvantages can be minimized by the use of more complex circuits.

Improved Closed-loop Regulator

Fig. 6 shows a widely used circuit which has relatively good performance. For analysis purposes, this is identical with the circuit of Fig. 5, the composite gain of the cathode-coupled stages being defined as $-M$.

The following improvements over Fig. 5 should be noted:—

(a) The reference tube is fed from the output rail and no longer has valve cathode current flowing in it, and thus tends to approach the performance of an ideal reference source. The high impedance loading of the grid circuit makes negligible current demand, and facilitates the use of a simple *RC* filter to remove the inherent tube noise.

(b) The use of cathode coupled amplifiers provides a measure of drift cancellation, and careful design can produce an improvement of about 4 : 1. Despite this improvement, contact potential drift is still the main factor in determining the stability of the output.

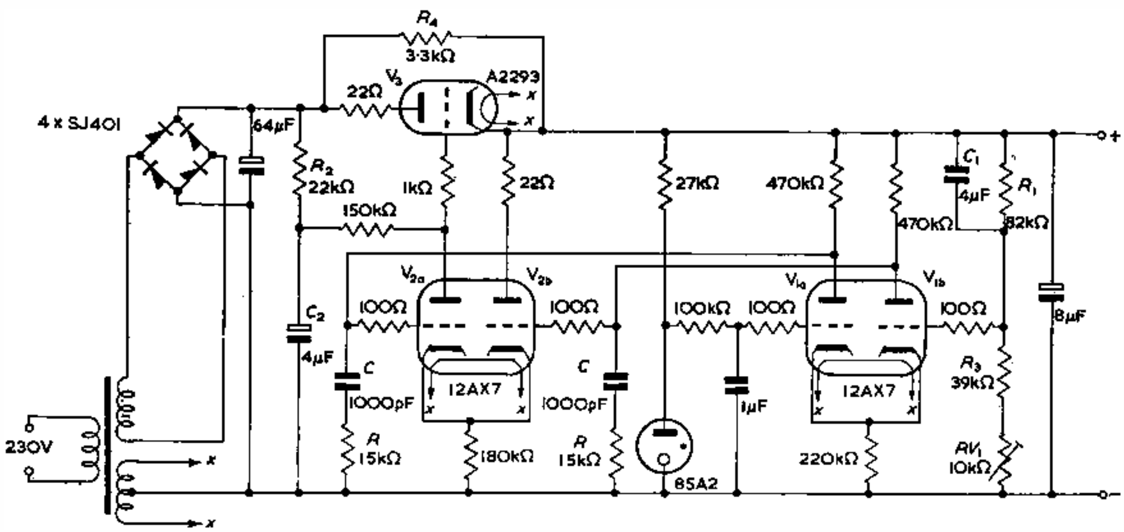


Fig. 7. Circuit diagram of experimental unit

The circuit of Fig. 5 serves to indicate some basic problems in the design of highly stable supplies.

It will be seen that the current feed for the reference tube is derived from the unstabilized supply, thus the reference voltage is susceptible to small variations due to changes in the unstabilized supply. This effect can be virtually eliminated by feeding the reference tube from the stabilized output.

A further disadvantage of the circuit of Fig. 5 is that the pentode current flows through the reference tube. Operational changes in this current will produce reference voltage changes by virtue of the dynamic resistance of the reference source. In addition, if the reference has a high dynamic resistance, the pentode gain is reduced.

Another problem is that of contact potential drift due to heater voltage variations. A measure of the problem is given by the fact that a 10 per cent variation in heater voltage can produce an equivalent grid-cathode signal of the order of 100mV. If one considers a supply of the type shown in Fig. 5 …

There are occasions when highly stable supplies are required and cannot be achieved by the above techniques. The designer must then resort to the use of chopper drift correction in the d.c. amplifier. This system is in common use in the instrumentation and analogue computing fields. When this system is used, a better reference is often also required.

There is available on the market a power supply[2] intended for analogue computing work, using a standard cell reference source and incorporating a drift corrected amplifier. This supply has an output voltage accuracy of 0·1 per cent, a long term stability of 0·1 per cent, and a stabilization factor of 4 000 : 1.

Design Considerations

Each design must be considered on its merits, and the resultant circuits should be no more complex than the specification requires. A power supply must have exceptional reliability, and thus must be of the simplest possible design. In order to illustrate one approach to a design problem, the Appendix contains design details of a simple fixed current load power supply.

General

By use of simple circuit configurations, some measure of improvement and/or simplification can be carried out. Typical examples shown in Fig. 7 are

(a) It is possible to increase the loop gain for hum and higher frequencies, by shunting the input arm of the feedback network with a capacitor. C_1 is used for this purpose.

(b) The R_2C_2 decoupling system reduces the breakthrough of the basic ripple into the amplifier loop.

(c) R_1, R_3 and RV_1 should be high stability type components. In high stability supplies, wirewound types are preferred.

(d) In systems which are not required to operate with zero load, the series valve complement can be reduced by the use of a 'bleed-by' resistor R_4. Its use in Fig. 7 is comparatively light. Care must be taken when using 'bleed-by', to ensure that the series valve is not cut off under conditions of high supply voltage and minimum load.

Loop Stability

A common problem to all closed loop stabilizers having high loop gain and more than two effective time-constants is that of loop stability.

Loop stability, in this sense, implies freedom from parasitic oscillations under all conditions of operation. By Nyquist theory, it can be shown that oscillation will inevitably occur at some frequency, unless precautions are taken to reduce loop gain without introducing excessive phase shift. If the loop gain can be reduced to unity and during the process adequate phase margin is maintained, the loop will be stable. The phase margin is a measure of the stability of the loop.

Safe reduction of loop gain can be effected by inserting a suitable combination of phase-shaping networks of the type shown in Fig. 8. This network has the property of introducing a known maximum attenuation and a phase characteristic which starts from zero, rises to a maximum, and finally reduces again to zero. By virtue of this, considerable attenuation can be introduced without the embarrassment of excessive phase shift. These techniques have been widely used to stabilize a variety of closed loop circuits, but every design must be treated as a separate problem with its own complexity.

A desirable property of a stabilized power unit is that it can be subjected to any degree of capacitive loading. Such capacitive loading can result from wiring capacitances and certain load configurations. In order to ensure absolute loop stability under these conditions, it is desirable to arrange the phase-shaping networks to cater for a purely capacitive load. In addition, in the interests of low source impedance over a large frequency range, a large capacitor should be built into the supply and connected across the output terminals.

Without this capacitor, the source impedance of the supply would increase as the loop gain decreased. With the capacitor, the impedance rises to a maximum at mid-band frequencies and falls away at the high frequency end of the spectrum.

The output capacitor makes the output time-constant large, and it becomes the major time-constant in the loop and limits the maximum phase margin to 90°, thus making the stabilization problem even more difficult.

Conclusion

Some aspects of power supply design have been discussed, but economy in design is being constantly improved by the introduction of new and more efficient components.

The circuit designer requiring power supplies is advised to tailor his circuits to work from standard packaged commercial units, wherever possible. If a special design is necessary, this article may indicate a method of approach.

Acknowledgments

The authors wish to record their thanks to the Directors of Solartron Research & Development Ltd for permission to publish this article.

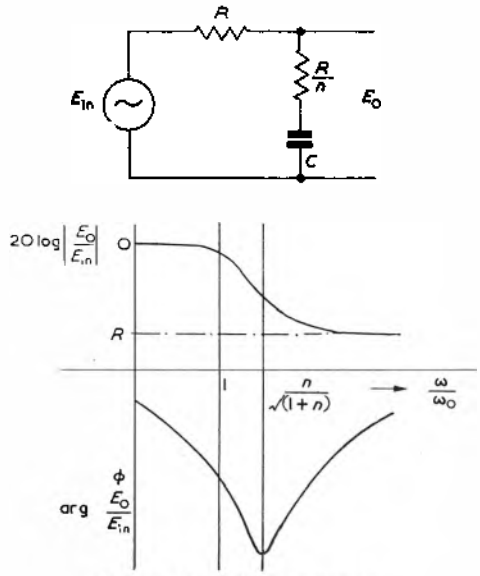


Fig. 8. Phase shaping network

Maximum ϕ occurs at $\omega/\omega_o = \dfrac{n}{\sqrt{(1+n)}}$

Maximum attenuation $R = 20 \log 1/1 + n$

Appendix

The design of a simple supply having the following specification is dealt with:

Output voltage	250V ±5V
Output current	50mA
Source impedance	Less than 1Ω, d.c. to 100kc/s
Noise and ripple	Less than 1mV peak-to-peak
Stabilization ratio	Greater than 400:1
Input voltage	Normal 230V, 50c/s ±5 per cent voltage variation

A circuit complying with this specification is shown in Fig. 7. The first step was to choose a suitable series element, and the choice fell on the G.E.C. A2293. For a 50mA current, a valve anode-cathode voltage of 30V is required at −1V bias. Thus the minimum reservoir voltage is 255 + 30 = 285V.

The current required for the d.c. amplifier and reference tube is supplied by a bleed resistor. This resistor allows 9mA to flow at minimum reservoir voltage.

The nominal reservoir voltage is 285V + 8 per cent. This makes allowance for transformer errors, giving a value of 308V, but a further allowance must be made for reservoir ripple. Using a 64μF reservoir capacitor and full wave rectification, the ripple was estimated at 10V peak-to-peak. The nominal d.c. level at the reservoir is thus $308 + 5 \simeq 315$V.

For economy, silicon junction rectifiers were used. Their high forward efficiency requires lower transformer voltage, thus enabling the designer to use electrolytic capacitors which can have high values for small physical size. A large reservoir capacitor eliminates the need for a choke, and thus the regulation is further improved. In order to utilize rectifiers of low p.i.v. rating, a bridge rectifier circuit was used.

Due to the use of a large reservoir capacitor, the peak charging currents are high, and this results in some clipping of the secondary waveform. The nominal 315V d.c. required would give an ideal secondary voltage of 230V r.m.s., but in practice a voltage of 240V r.m.s. is required. The peak a.c. voltage which is encountered at the highest secondary voltage condition is 340V.

On consideration of source impedance and ripple, a loop gain, βM, of at least 60dB was required in the amplifier.

A convenient reference tube is the Mullard 85A2 which gives a feedback network transmission, β, of approximately one-third at d.c. This implied an amplifier gain, M, of >70dB. Use of two 12AX7 cascaded amplifiers, under the conditions shown, produces the required gain.

Care has been taken during the design to cater for the voltage variations which occur due to loop operation and due to component tolerances. A point that must be noted is that the bleed current under the worst conditions can be approximately 30mA, the valve current thus being reduced to 30mA. This requires a grid swing of approximately 19V.

The stabilization ratio required is met easily by a cathode-coupled amplifier.

In order to produce low source impedance at high frequencies, the output terminals are shunted with a large value capacitor. This is essential, due to the difficulties of maintaining the high loop gain at the higher frequencies, but its introduction makes the output time-constant large. This large time-constant is the major time-constant in the loop. For the present design, this time-constant has a 3dB frequency of about 200c/s, and has introduced a phase shift of greater than 86° at frequencies above 2kc/s. The resultant attenuation at 2kc/s is 20dB, and is increasing at the rate of 20dB per decade.

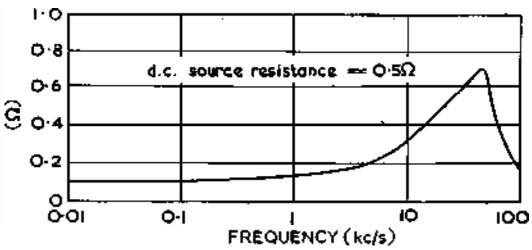


Fig. 9. Source impedance of experimental unit 5Ω

There are two other significant time-constants in the loop, those associated with V_{1a} anode and V_{2a} anode. The time-constant at V_{1a} anode was estimated to have a 3dB frequency limit of about 10kc/s, and that at V_{2a} anode a limit of 55kc/s. Due to these additional time-constants, there would be instability between these frequency limits. In order to prevent this instability, it is necessary to reduce the loop gain at the critical part of the spectrum. This is achieved by use of a phase shaping network, and it is convenient to modify the anode time-constants of V_1 by means of the networks, RC. Using the values shown, V_{1a} time-constant has a new 3dB frequency of 1kc/s, and at 55kc/s has introduced an attenuation of approximately 21dB with a phase shift of about 10°. The network introduces a maximum phase shift of about 57° at 3kc/s, but this is of no consequence, since the total phase shift in the loop at this frequency is then only 148°.

The loop gain has reduced to unity by about 60kc/s, and up to this frequency the phase margin has always been greater than 30°.

The measured performance figures on an experimental unit were:

Output voltage range	240 to 260V at 50mA
Source impedance	as plotted on Fig. 9
Noise and ripple	0·8mV peak-to-peak at 50mA
Stabilization Ratio for ±5 per cent	800 : 1

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A High Stability Electrostatic Generator

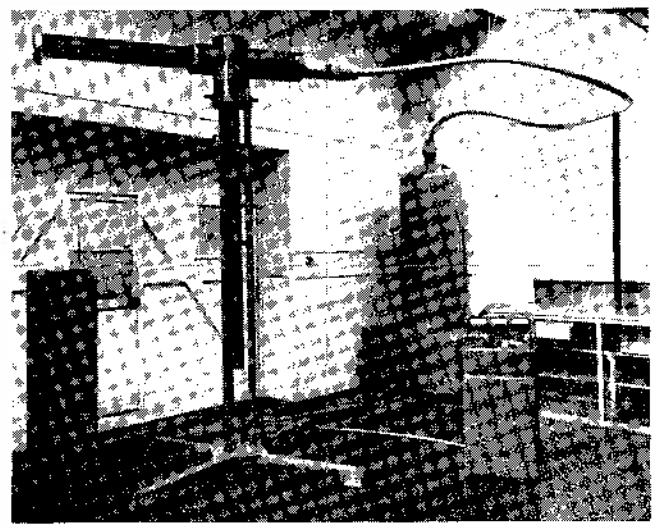
A high stability electrostatic generator was recently demonstrated by Miles Hivolt at their Shoreham Airport works.

This machine which will deliver an output of 600kV at 4mA is made by the Societe Anonyme de Machines Electrostatiques (SAMES) of Grenoble and consists essentially of a vertical cylinder pressurized to 15-25 atmospheres and containing a rotor of insulating material driven at 3 000rev/min by a motor at the base of the cylinder.

Excitation of the rotor is provided by an r.f. oscillator and rectifier equipment and the output voltage, which is adjustable to any value up to 600kV, can be stabilized to 0·1 per cent.

A separate control unit contains protective circuits against voltage and current overloads and an additional feature is a circuit designed to lower the output voltage when sparking occurs and to raise it slowly again to its former value when sparking has ceased.

Because of its small size and low weight it is readily transportable and is thus suited for field tests of transmission lines. The generator may also be used on a high voltage accelerator.


The SAMES 600kV generator with its associated control unit and power cabinet.

Telemetering Employing the Principle of the Gas-filled Stepping Tube

By Yoshisuke Hatta*

A method of telemetering of mechanical displacement employing the principle of the gas-filled stepping tube is described. As it is an electronic device it has good high speed operation, and is, therefore, superior in some respects to electromagnetic systems which have considerable inertia.

(Voir page 248 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 254)

TELEMETERING of mechanical displacement is usually achieved by using the self-synchronous motor but, because of its inertia it is incapable of the quick response which modern telemetering demands. This article presents a new method of telemetering using the principle of the gas-filled stepping tube. As this is an electronic device its speed of response is excellent and, moreover, it is convenient for practical applications because the whole device is very simple.

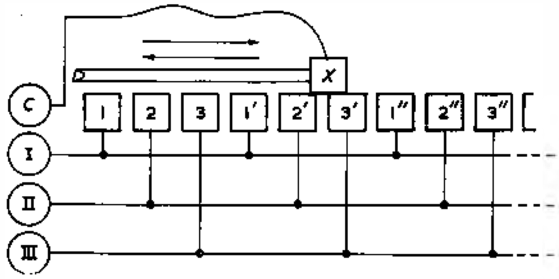


Fig. 1. Transmitter of 3 channel method

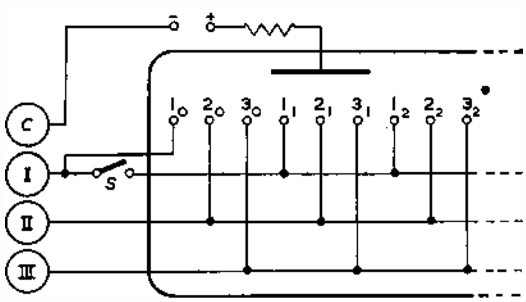


Fig. 2. Indicator (or receptor) of 3 channel method

Three Channel Method

The principle of this method is shown in Figs. 1 and 2. Fig. 1 represents the elementary structure of the transmitter which converts the mechanical displacement to the electric signal. It has contacts, 1, 2, 3, 1′, 2′, 3′, ..., and a sliding contact (X) which follows the displacement to be measured. Fig. 2 represents the indicator (or receptor) which consists of a gas-filled stepping tube. It has a common anode (A) and rod like cathodes in three groups, (1_0, 2_0, 3_0, ...) (1_1, 2_1, 3_1, ...) (1_2, 2_2, 3_2, ...). Each group is called 1, 2 and 3 successively. The transmitter and the receptor are connected by 4 lines (I), (II), (III) and C which form three channels between them. In normal operation the switch (S) is closed. Now, take a case in which X is on the contact 2′, and the glow of the stepping tube is at 2_1. Then, as X moves from 2′ to 3′, the glow transfers from 2_1 to 3_1 which is closest to 2_1 among the cathodes of 3 group, according to the principle of the gas-filled stepping tube. If X moves reversely from 2′ to 1′, the glow transfers from 2_1 to 1_1 for the same reason. Thus, one to one correspondence between the transmitter and the receptor is secured independently of the number of contacts, and the glow follows the movement of X. S is used for resetting. If it is required to reset the glow to 1_0, X is put on 1 and S is opened when the glow will jump to 1_0 from any position of 1 group. After resetting, S is closed and the whole set enters into normal working.

The stepping tube shown in Fig. 2 is the same as the double pulse Dekatron in structure, but differs from it in

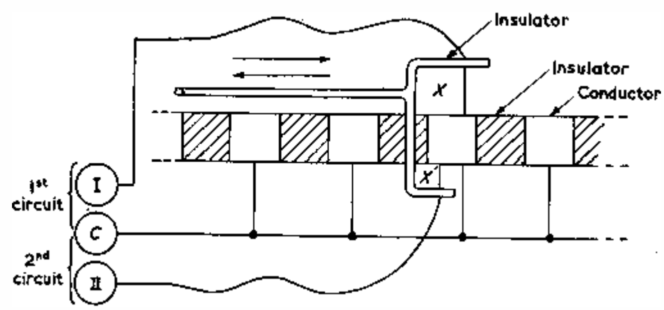


Fig. 3. Transmitter of 2 channel method

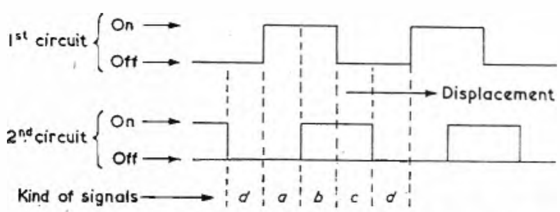


Fig. 4. Output signals of the transmitter shown in Fig. 3

mode of operation; i.e. while the cathodes of 1 group operate as index cathodes and the cathodes of 2 and 3 groups are used as guides in the Dekatron, all cathodes act as index cathodes in the stepping tube.

Two Channel Method

Figs. 3 and 4 show the elementary structure and the waveforms of the output signals of the transmitter for the two channel method. These were originally used by Kondo and Takahashi in the so-called K.T. wheel[1]. It will be seen that the waveforms of Fig. 4 are similar to those of the input voltages of the double pulse Dekatron. So, when those output signals are applied to a double pulse Dekatron of conventional type, with appropriate circuit constants, the glowing position will move following the movement of X. But the double pulse Dekatron is inadequate as an indicator tube because it has cathodes of three groups while the transmitter shown in Fig. 3 generates four kinds of

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signals per cycle and, consequently there is a rather complex relationship between the displacement of X and the glowing position.

Due to these considerations, the author makes use of a stepping tube which has cathodes in 4 groups (the single pulse Dekatron has an electrode structure of this type).

The operation of the circuit shown in Fig. 5 can be explained as follows. Three terminals marked (I), (II) and C are connected to the corresponding terminals in Fig. 3 by 3 lines forming 2 channels between the transmitter and the receptor. First, consider the operation of V_1 (double triode). When the first circuit is open, V_{1a} is cut off by the potential drop across R_2, while V_{1b} is conducting because its grid is positively biased by R_3 and R_g. But when the first circuit is closed, E_b over-compensates the negative bias on V_{1a} given by R_2 and it becomes conducting, while

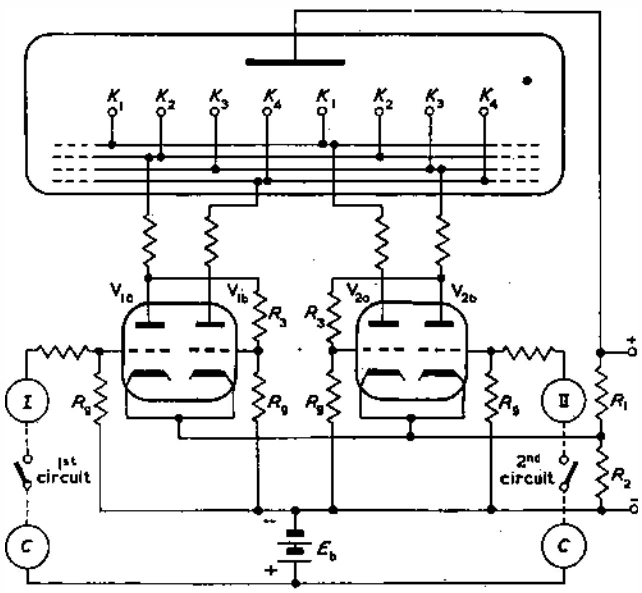

Fig. 5. **Receptor circuit of 2 channel method**

V_{1b} is cut off because the potential of its grid falls in accordance with the anode potential of V_{1a}. The operation of V_2 is similar to that of V_1. Owing to the operation of V_1 and V_2, two adjacent cathodes (K_1, K_2) will glow corresponding to the signal a of Fig. 4 and in the same way (K_2, K_3), (K_3, K_4) and (K_4, K_1) will glow corresponding to the signals b, c and d of the same figure, successively. So in the indicator tube, two adjacent cathodes are always glowing simultaneously and the position of glow follows the displacement X of Fig. 3 with linear proportionality. This mode of operation is quite different from that of conventional types of Dekatrons.

Some Developments

For measuring rotational displacement, the transmitters for linear displacement must be modified as shown in Figs. 6 and 7. In rotational displacement, the number of rotations may increase indefinitely. When this is the case, the receptor must have a decimal counting circuit which is capable of indicating the number of rotations. This decimal counting circuit which consists of a number of Dekatrons must have reversible coupling circuits[2] between two successive orders instead of the uni-directional ones used in conventional counting circuits.

As an example of the application of the three channel method, the author applied this principle to the indication of the reading of a dial-gauge. The transmitter which had

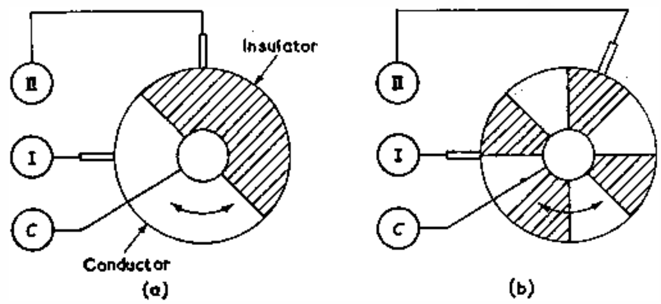

Fig. 6. **Transmitters of 3 channel method for rotational displacement**
(a) 1 cycle per rotation
(b) 4 cycles per rotation

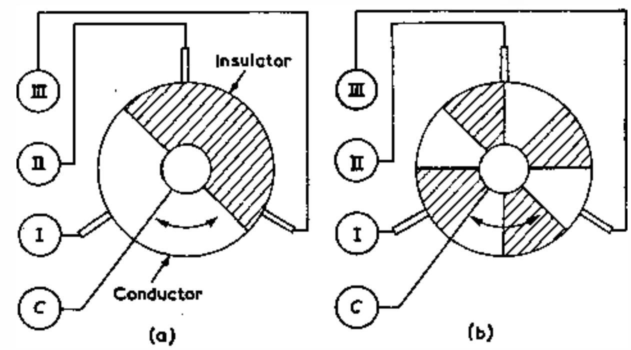

Fig. 7. **Transmitters of 2 channel method for rotational displacement**
(a) 1 cycle per rotation
(b) 4 cycles per rotation

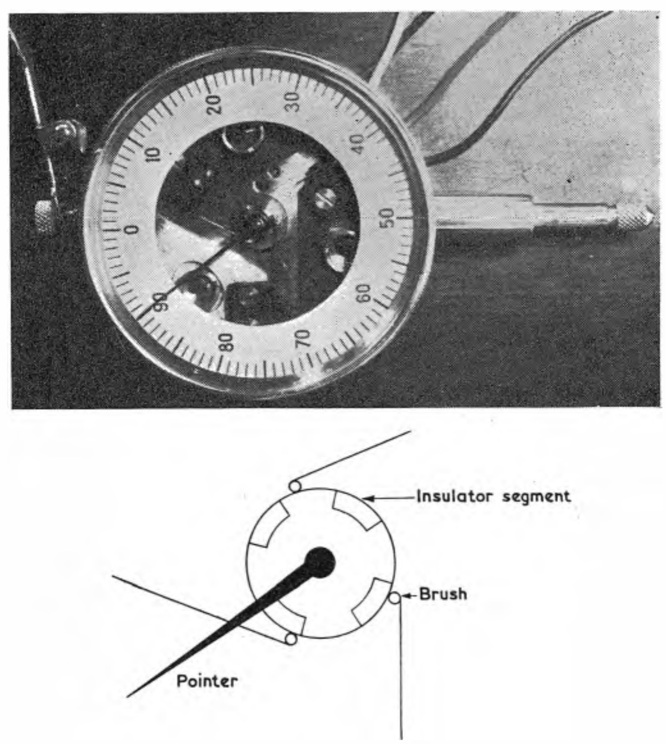

Fig. 8. **Dial-gauge with the transmitter applied to the centre axis**

the structure shown in Fig. 6 was applied to the main axis of a dial-gauge (Fig. 8). As an indicator tube, the author designed a linear form stepping tube which had 121 cathodes and could cover 10 rotations of the pointer of the dial-gauge.

The operation of the whole equipment was almost satisfactory. But, when the speed of rotation was high, the output signals of the transmitter were sometimes erratic which was due to jumping of brushes and seemed to be inevitable so long as mechanical contacts were used. After replacing the mechanical switching element by photo-electric ones employing a light beam, a photo-transistor and a rotating shutter, the reliability of operation was improved.

Conclusions

The merits of this method of telemetering compared with others having similar functions are:

(1) The indicator is capable of giving sufficiently quick response for the measurement of mechanical displacement, and there is no slip between the displacement to be measured and the indication of the receptor.

(2) As it is a digital device, its operation is insensitive to drifts of circuit constants and fluctuations of supply voltage.

(3) A small and frictionless transmitter can be designed so as to give no disturbance to the movement of displacement to be measured.

Acknowledgments

The author wishes to express his thanks to Prof. Y. Watanabe who directed the work, to Prof. M. Kondo for his collaboration and to Prof. H. Takahashi for his valuable suggestions. This work was supported by a grant in aid for Miscellaneous Scientific Research of Japanese Government.

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A Three Digit Voltmeter

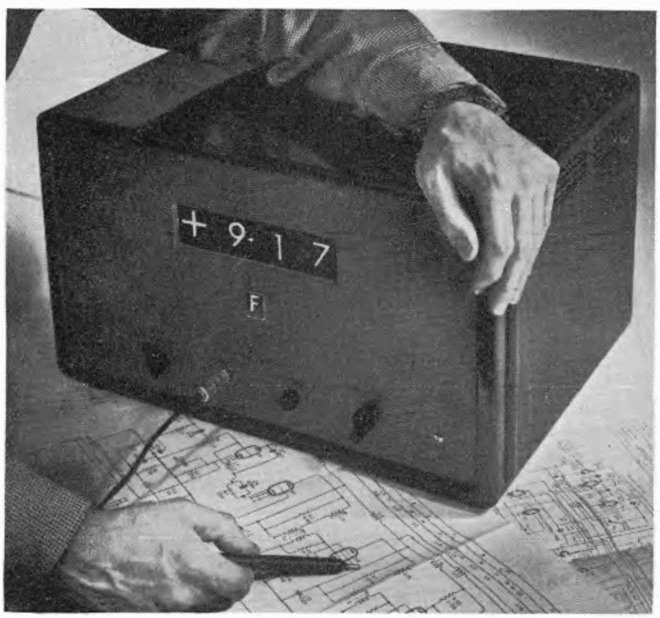
Ferranti Ltd have announced the introduction of their three-digit voltmeter D101 which incorporates full automatic ranging and polarity selection, the three ranges being as follows:

0·01 to 9·99V d.c. 10·0 to 99·9V d.c. 100 to 999V d.c.

The appropriate range is automatically selected to give the most significant reading so that, for example, 9·76V cannot be displayed as 09·8V or 009V. The resolution and linearity of the instrument is 0·1 per cent of the full scale reading on any range.

The measured voltage is read directly in digital form on a $1\frac{1}{4}$in by $5\frac{1}{2}$in projection display until giving 1in high numerals. The polarity and decimal point are also displayed, the decimal point placement being determined by the range selected. While the voltmeter reads only d.c., an a.c. to d.c. convertor is under development. The unique logic employed in the switching circuits considerably reduces the reading time which is 700msec on average, including ranging. The instrument has been designed for use on either 50c/s or 60c/s mains, and transistorized circuits give freedom from drift due to mains variations. High grade choppers have been incorporated to give complete stability and printed circuits have been used whenever possible to achieve the utmost reliability. To ensure the absolute accuracy of the instrument, a bank of nine Weston standard cells is built in for calibration. A suitable printer is under development for coupling directly to the voltmeter, all relevant information being presented at a socket at the rear of the unit, and it will be capable of recording two or three readings per second. A remote display may also be fitted instead of or in addition to the internal readout. The instrument is contained in a bench type cabinet measuring 17in by 13in by $10\frac{1}{2}$in high, the weight being 50lb approximately.

The digital voltmeter



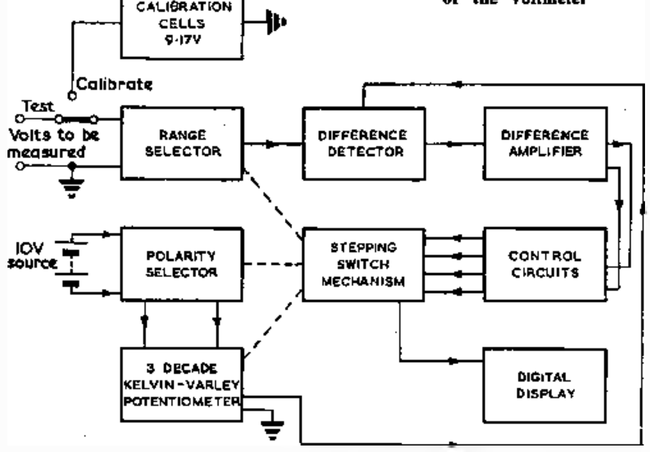


Fig. 1. General arrangement of the voltmeter

The principles of operation may be considered by referring to Fig. 1. The applied d.c. voltage is compared with the output from an internal decade Kelvin Varley potentiometer fed from a 10V d.c. reference source. Any difference between the two voltages is detected and causes the stepping switches to operate to select the correct range and polarity, and then to step the decades of the potentiometer to a 'balance', the latter being achieved when the difference between the two voltages is less than 6mV. The mercury cell reference source is calibrated against the bank of standard cells giving 1V, 100mV and 10mV steps respectively on the three decades of the Kelvin Varley network, and the average cell life is 1 000 hours. A 'stand-by' position is incorporated on the mains on-off switch, which allows the cells to be disconnected when the instrument is not actually reading. These cells are positioned to afford ease of access, being located at the rear of the unit, and an alternative reference source may be fed in if the instrument is to be used as a ratiometer. The Weston standard cell bank is also easily removable to enable it to be used elsewhere; alternatively, an external calibration source may be connected.

The limitations of moving pointer voltmeters are overcome by digital instruments, which are inherently more accurate, and also eliminate the fatigue associated with continuous reading. Moreover, it is possible to use less highly skilled labour in this task, thus releasing technical personnel for more important duties. Applications for the three digit voltmeter exist in the fields of automatic testing and monitoring, and wherever information can be presented in the form of a voltage. Examples of its use in production control are voltage measurement of batteries and for measuring thickness of materials by using voltages as parameters.

Solid Electrolyte Tantalum Capacitors

By Prof. Robert Aries*

Solid electrolyte tantalum capacitors have found ready acceptance for use in miniaturized equipment. In this article a brief description of the manufacturing process is given together with the results of life tests and temperature characteristics.

(Voir page 248 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 254)

WHILE semiconductors are well suited for miniaturized equipment, the associated components must also be miniature and have properties to match the high performances of the new transistors and rectifiers. An advance in this direction has been the successful manufacture on a large scale of miniature tantalum capacitors with a solid electrolyte.

While a non-liquid electrolyte capacitor was proposed more than 30 years ago, and has been known on a laboratory scale for two decades, the more recent emergence of better techniques, coupled with a strong demand for its performance characteristics has made the solid tantalum capacitor an industrial reality.

The use of solid electrolyte capacitors has been increased rapidly because of their obvious advantages. Electrolyte evaporation ceases to be a problem, the temperature range drops to −80°C, true hermetic sealing is feasible, and shelf life is indefinite.

While a number of solid electrolytes or semiconductors are known and can theoretically form suitable semiconductors between the dielectric and the cathode, and thus may be considered part of the cathode, there is the real difficulty that it is no easy matter to provide the proper related physical distribution in space between the dielectric and the semiconductor 'solid electrolyte'. One solution to this problem is given by Van Geel in which manganese dioxide is indicated as a semiconductor. In one varient a concentrated manganous nitrate solution is imbibed into the pores of an anodized sintered tantalum powder pellet or slug, and then the manganous nitrate is decomposed at elevated temperatures to form manganese dioxide. This is a violent reaction, liberating extensive volumes of reactive nitrogen oxide gases, and a great deal of art and experience is required to achieve successful results. Damage occurring fortuitously to the dielectric film must later be healed, and a series of anodizings to form and to heal the dielectric and to coat it can be carried out, which makes the entire operation both an art and a technology. When these operations are skilfully carried out the resulting unit, which is ultimately hermetically encapsulated, provides a unit small in size, high in capacitance, stable, trouble-free in operation, and with a practically unlimited life.

Although the recommended maximum operating temperature is 85°C, investigations indicate that use at temperatures up to 125°C is satisfactory, and particularly so if the voltages used are a little lower than the rated voltages. The voltage ratings available in solid electrolyte capacitors are 6V to 60V, but work is in hand to extend the upper limit.

Investigations with units from the 6V to the 60V range have shown that the capacitance is almost linear with temperature and that the capacitance increases with rising temperature. Thus, at 85°C the capacitance is 105 per cent of the capacitance at 20°C, so that the percentage coefficient of rise of capacitance per degree is 5/65 or 0·077 per cent/°C for the 60V units. This is shown in Fig. 1. A capacitance of up to 400μF has already been achieved in these units.

With lower voltage higher capacitance capacitors, the percentage coefficients are somewhat lower, being about 0·06 per cent/°C for the 6V units. The outstanding characteristic is that capacitance within the operational range increases somewhat between room temperature and the currently permissible highest operating temperatures by an

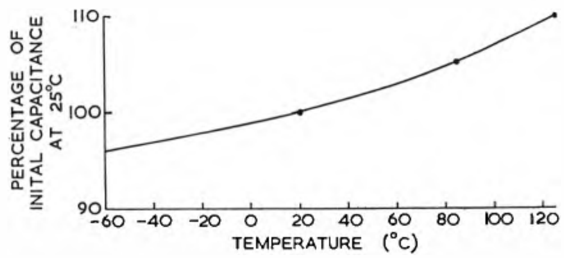


Fig. 1. Change of capacitance with temperature

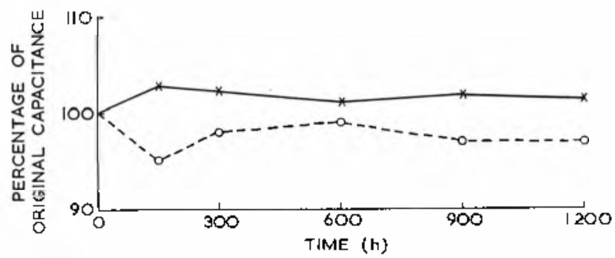


Fig. 2. Results of life tests

- - - - - - *per cent of original capacitance at 85°C measured after hours indicated at 85°C*

——— *per cent of original capacitance at 25°C measured after hours indicated at 85°C*

amount between 5 and 10 per cent. Such units, after life testing for 1 000 hours at 60°C, show no appreciable change from the original room-temperature characteristics when again tested at room temperature.

A large number of 60V capacitors (selected to give substantially equal initial capacitances to begin with) have been tested under various conditions, by removing random samples after life tests of 300, 600, 900 and 1 200 hours at 85°C. The capacitance was determined in each case on separate specimens at 85°C. The results are shown in Fig. 2.

This indicates that on life testing at the currently extreme temperature of 85°C for up to 1 200 hours there is a slight increase in the capacitance at 25°C (about 1 per cent) and a slight decrease in the capacitance at 85°C (not exceeding 4 per cent). Taking into account the rise in capacitance with increasing temperature (about 5 per cent at 85°C) the capacitance at 25°C and at 85°C after 1 200 hours of life testing at the rated voltage,

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remains slightly higher than the initial rated capacitance at 25°C.

Leakage current is, of course, an important characteristic of capacitors subject to specification and the leakage current rises rather sharply with temperature, but is almost independent of the voltage rating and the rated capacitance, and is possibly a characteristic of the geometry of the capacitor and the temperature for this type of capacitor.

A 60V unit of 0·6μF capacitance showed the following leakage losses at various temperatures:

Temperature (°C)	Leakage (μA)
0	0·11
20	0·15
25	0·23
40	0·41
50	0·57
60	0·80
75	1·20
85	1·75

This indicates that a rise in temperature of 20 to 25°C about doubles the leakage loss for the range of 0 to 85°C. This is shown in Fig. 3.

Alongside many other new products in electronics, the production of tantalum capacitors is a new and interesting development. Continued improvements will be made in it, so that it is not unreasonable to expect increased capacitance, voltage ratings and temperature ranges.

While research under way may eventually change the metal to niobium or zirconium, the solid type may be considered as important a development as electrolytics were at their inception.

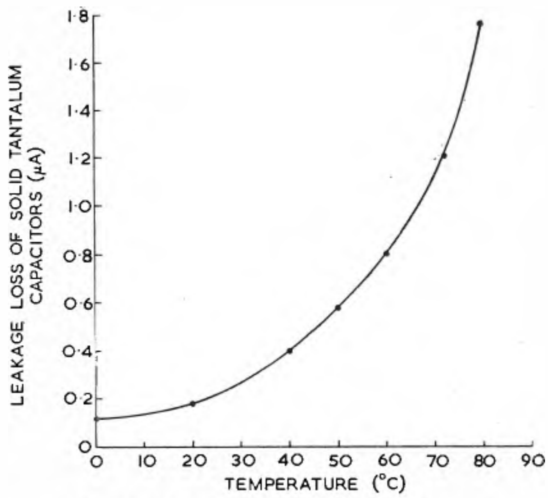


Fig. 3. Variation of leakage with temperature

New High Power Transmitters at Ongar

Seven new transmitters have recently been brought into operation at the G.P.O. radio station at Ongar, Essex.

Designed to a Post Office specification, they were manufactured by Standard Telephones and Cables Ltd and they incorporate the latest developments in high-power, high-frequency transmitter design. To provide for maximum flexibility in use, the transmitter equipment is divided into two groups. The first group is arranged to receive the signals sent by line from the sending office, to convert them to a standard frequency (3·1Mc/s) and to deliver them to the second group—transmitter proper—at a standard power level (0·25W). The transmitters can each radiate a peak power of 30kW at any frequency between 4 and 27·5Mc/s and remote control facilities allow any transmitter to be run up on any one of six pre-selected frequencies, with ability to change from one frequency to another at will—the change takes less than a minute. Similarly, the transmitters can be closed down by remote control.

Facilities are provided for the monitoring of the signals leaving the transmitters, and a pick-up from the output of any transmitter can be fed back to a control room for inspection by a spectrum analyser and other monitoring apparatus to ensure correct operation.

Elaborate steps have been taken to ensure safety, both of the staff, and for protection of the equipment itself. All the oil-filled apparatus is enclosed in a fire-proof room, and the transmitter cubicles are fully interlocked with the main power supply switch, so that access to the interior of the equipment cannot be obtained until the power is disconnected. The power supply is taken from the public mains at 11kV and transformed down at each station to feed the equipment at medium voltage. Standby diesel alternators can replace the public supply in the event of a failure.

This latest addition to the capacity of the Ongar Radio Station will enable the Post Office to provide more and better international telegraph services to most parts of the world. These services include international telex, picture telegraphy and the leasing of circuits to customers for their individual use. There are three separate transmitter buildings on the station operating four long-wave and thirty-three short-wave transmitters. Two of the three buildings are normally unstaffed, and the transmitters in them are remotely operated from the third.

One interesting feature of the Station is the 300ft masts carrying the long-wave aerials that service Europe, and the short-wave Franklin aerials for the services to Bombay, Melbourne, New York and other overseas territories. The new transmitters use rhombic aerials, mounted in pairs one above the other on the same masts, and employing different frequencies, an aerial switch automatically connecting the correct aerial according to the wavelength in use at the time.

The first radio station was established on the site by the Marconi Wireless Telegraphy Company in 1930. Subsequently developed by Imperial and International Communications (later Cable and Wireless), the station was taken over by the Post Office in 1950.

In its early days the Station used long-waves for its services to the Continent, but by 1931 short-wave transmitters had been installed and the Station had become Britain's principal radio station in the beam telegraph services to the Commonwealth.

LETTERS TO THE EDITOR

(We do not hold ourselves responsible for the opinions of our correspondents)

Response of a Capacitance-Resistance Divider

DEAR SIR,—Mr. S. Turk's useful results in your October issue for three specific forms of input may be generalized for an arbitrary input which may be known only numerically or as an oscillogram. If the divider has zero initial stored energy and $V_s(t)$ is the response to a unit step function input, then by Duhamel's Theorem the response to any input $V_1(t)$ is given by

$$V_2(t) = V_s(o)\,V_1(t) + \int_0^t V_1(x)\,V_s'(t-x)\,dx \qquad (1)$$

where V_s' denotes the first derivative. Substituting the known response

$$V_s(t) = k\,(1 - \alpha e^{-t/\nu})$$

where

$$k = \frac{R_2}{R_1 + R_2}\ ;\ \nu = \frac{R_1R_2\,(C_1 + C_2)}{R_1 + R_2}\ ;$$

$$\alpha = 1 - a = 1 - \frac{(R_1+R_2)\,C_1}{R_2\,(C_1+C_2)}$$

one obtains

$$\frac{V_2(t)}{k} = (1-\alpha)\,V_1(t) + \alpha/\nu \int_0^t V_1(x)\,e^{-(t-x)/\nu}\,dx \qquad (2)$$

It is convenient to introduce what may be termed the *distortion* of the divider, a function of time defined by

$$D(t) = V_1(t) - V_2(t)/k \qquad (3)$$

This function indicates the extent to which the response to a particular form of input deviates from that of an ideal divider having a constant division ratio k, and it is of course zero for a pure d.c. input. Substituting equation (2) in equation (3) one obtains

$$D(t) = \alpha\left\{V_1(t) - 1/\nu\int_0^t V_1(x)\,e^{-(t-x)/\nu}\,dx\right\} \qquad (4)$$

This expression may be evaluated numerically if $V_1(t)$ is only known numerically. It is immediately apparent that for all forms of input the distortion is zero if $C_1R_1 = C_2R_2$, since in this case $\alpha = 0$.

By successive integrations by parts, the following series form of the expression is obtained.

$$D(t)/\alpha = \nu V_1'(t) - \nu^2 V_1''(t) - \nu^3 V_1'''(t) \ldots + e^{-t/\nu}\left\{V_1(0) - \nu V_1'(0) + \nu^2 V_1''(0)\right\} \qquad (5)$$

The main application of the series is for the case of a small value of the time-constant ν, or for a slowly varying input, where the region of interest is not too near the origin (say $t > 4\nu$). Then the second series, which represents the initial transient, is negligible and the first series converges rapidly. A graphical method

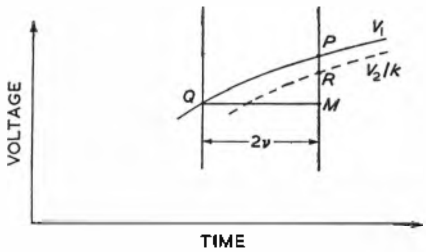


Fig. 1. Graphical method for determining distortion

for determining the distortion in this case is shown in Fig. 1, where P is the point on the curve of V_1 corresponding to time t. Draw a vertical through P and another a horizontal distance 2ν to the left, cutting the curve at Q. Draw a horizontal through Q cutting PM at M. Then it may be shown that $\alpha/2 \times PM$ is the distortion at time t, with an error of about one third of the third term of the series. Hence if $PR = \alpha/2 \times PM$, then R is a point on the response curve V_2/k, and by repeating the process a portion of the response curve may quickly be drawn.

A limitation of the series form (5) is that it is applicable only in an interval in which there are no discontinuities of V_1 or its derivatives other than at the origin. For example, it cannot be applied to the ramp function of Fig. 7 of the article. If there are discontinuities at the origin, each term is given the value it takes just *after* the discontinuity.

When the input is known as an algebraic function, an advantage of expressing the response in terms of the distortion is that it usually involves fewer independent variables, and hence may be represented by a smaller family of curves. For example, the distortion for a step function input is, in the notation of the article, given by

$$D/A_0\,(1-a) = e^{-t/\nu} \qquad (6)$$

and for an exponential input

$$D/A_0\,(1-a) = \frac{\Gamma}{1-\Gamma}\left\{e^{-\Gamma t/\nu} - e^{-t/\nu}\right\} \qquad (7)$$

These equations are simpler than equations (3) and (5) of the article. If preferred, the *fractional* distortion may be plotted, without increasing the number of variables, by dividing by the input function V_1. For all forms of input, the fractional distortion at $t = 0$ is $(1 - a)$.

Yours faithfully,

R. H. EVANS

Farnborough, Hants.

The Author Replies:

DEAR SIR,—Mr. Evans is treating the divided response in a more general way. However, his graphical method applies mainly to the later part of the response where the transients are disappearing or negligible, while I put emphasis on the initial part of the waveform. I therefore consider his results as a generalization in range, that was not of main interest in my article. This very point Mr. Evans made clear in his letter.

Yours faithfully,

S. TURK,

Institut 'Ruder Boskovic,' Zagreb, Yugoslavia.

A Cordless Telephone Switchboard

DEAR SIR,—In your January issue you published details of a Cordless Telephone Switchboard installed at the European headquarters of the Canadian Pacific Railway Company in London.

The first cordless board in this country was installed by Siemens Bros Ltd (now Siemens Edison Swan) for the Post Office at Thanet in 1955. Reference to this can be found in the *Post Office Engineering Journal*, Vol. 48, Part 2, July 1955. Other cordless exchanges have been under construction since this date, including one at the Royal Victoria Hospital, Belfast, which has been working for the past year.

Yours faithfully,

N. F. SAY

Streatham, London, S.W.2.

'Droop' and Low Frequency Bandwidth

DEAR SIR,—While sufficiently accurate relationships between the rise-time of a pulse and the -3dB high frequency bandwidth of an amplifier are well known, it would appear that the relationship between the 'droop' of a pulse and the low frequency bandwidth is less well known.

For small droops, of the order of 15 per cent, such as are chiefly of interest in amplifier design, the relationship may be stated as:—

$$D = \omega\tau \qquad (1)$$

where D = Droop

$$= \frac{e_o - e_{o_\tau}}{e_o}$$

e_o = output voltage at $t = 0$

e_{o_τ} = output voltage at $t = \tau$
$\omega = 2\pi f_1$
$f_1 = -3$dB low frequency band edge.
τ = pulse duration.
$\alpha = 1/CR$

This may be shown as follows:—

The droop, as defined, is given by

$$D = \frac{e_0 - e_{0_\tau}}{e_0} = 1 - e^{-\alpha\tau} \quad \ldots\ldots\ldots\ldots\ldots (2)$$

The low frequency band edge is defined by

$$\omega = 1/CR \quad \ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots (3)$$

For small values D, less than 0·15, or 15 per cent sag,

$$D = \alpha\tau \quad \ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots (4)$$

Substituting equation (3) into equation (4), observing that $\alpha = \omega$.

$$D = \omega\tau \quad \ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots (1)$$

Whence $f_1 = \dfrac{D}{2\pi\tau}$ etc.

Accurately

$$f_1 = \frac{\ln 1/(1 - D)}{2\pi\tau} \quad \ldots\ldots\ldots\ldots (5)$$

It will be agreed that equation (1), a particularly easy expression to remember and use, is of more practical use than equation (5) or some other published equations invoking a knowledge of the amplifier phase response.

In the case of N non-interacting couplings it will be found that equation (1), for the purpose of quick estimation, gives a pessimistic answer.

Yours faithfully,
T. G. Clark
Radar Development Laboratory,
Decca Radar Ltd.

Design Curves for Simple Filters

Dear Sir,—With reference to the article by D. J. H. Maclean of Messrs. Barr and Stroud, in the November issue of Electronic Engineering, I would be grateful if you would pass the following query on to the author.

In Example 3 in the above article there are two tables, one specifying a band-pass filter of 50kc/s bandwidth and the other a low-pass filter from which the first is derived.

I suggest that the second table should read:—

Attenuation	*Frequency Interval*	*Normalized Interval*
$\leqslant$0·5dB	$0 \leqslant f \leqslant 50$kc/s	$0 \leqslant \omega/\omega\beta \leqslant 1$
$\geqslant$ 50dB	$75 \leqslant f \leqslant \infty$	$1{\cdot}5 \leqslant \omega/\omega\beta \leqslant \infty$

This is consistent with the interval from the assumed zero to the upper 50dB point being 75kc/s. If, as the author states, it is the total interval between the 50dB points that must be considered, i.e. 100kc/s, why, in the table in Example 1, is the frequency interval given as 2kc/s and not 3kc/s, when 3kc/s is the total frequency interval between the 36dB points.

There seems to be an inconsistency between these two tables, both of which make use of the same assumptions in specifying low-pass filters.

Yours faithfully,
L. B. Jones
Bristol Aero-Engines Ltd,
Bristol.

The Author Replies:

Dear Sir,—The apparent inconsistency between Example 1 and Example 3 which Mr. Jones has pointed out is, I suggest, based on a misunderstanding. Example 1 deals with a straightforward low-pass filter, consequently there is only one 36dB point on the attenuation characteristic, not two as Mr. Jones implies. In the case of Example 3, the low-pass/band-pass transformation is involved, an important property of which is that when going from a coil to a series resonant circuit (and from a capacitor to a parallel resonant circuit), the bandwidth in cycles at any given attenuation level remains the same, if the coil (and capacitor) is unchanged (see, for example, 'Network Analysis and Feedback Amplifier Design', pp. 211-214, by H. W. Bode). As one can see from the Example 3, this is exactly what is done, and therefore the given table for the low-pass prototype is correct.

Yours faithfully,
Douglas J. H. Maclean,
Barr and Stroud Ltd,
Glasgow.

Characteristics of Transistors

Dear Sir,—In the course of investigating the V_c-I_c characteristics, at high currents, of a number of types of junction transistor a second regression in the avalanche characteristic has been observed which appears to have received little, if any, attention in published literature.

Typical characteristics are shown in the accompanying figures reproduced from oscilloscope trace photographs; the scanning current was a single-shot ramp pulse of 200μsec duration. Traces shown in the first three figures are:

(1) V_{cb} characteristic with emitter open-circuited.
(2) V_{cb} or V_{ce} characteristic with base shorted to emitter.
(3) V_{ce} characteristic with base open-circuited.

Fig. 1(a), (b) and (c) are for samples of a medium power junction transistor, the form of the regression being clearly seen as a rapid voltage break-back occurring at 2·6A in trace (b) and at 1·7A in trace (c) (curves (2)). Two interesting features appear: firstly that the effect occurs most readily in the base-emitter short-circuit condition, secondly that the effect is incipient in the collector-base characteristic (1) of Fig. 1(c) at a current approximately the same as that at which breakdown occurs in the base emitter short-circuit conditions. Trace (1) in (c) was made before trace (2).

Tracing the reverse characteristics of a number of germanium junction diodes a somewhat similar breakdown appears, a trace for one sample being reproduced in Fig. 1(d). For both transistors and diodes scanning the characteristic results, in some instances, in the destruction of the device; in other instances the process of scanning the characteristic beyond the regression point may apparently be repeated indefinitely without deterioration of the device, the position on the characteristic at which the break-back occurs being apparently not related to the rated power dissipation of the device.

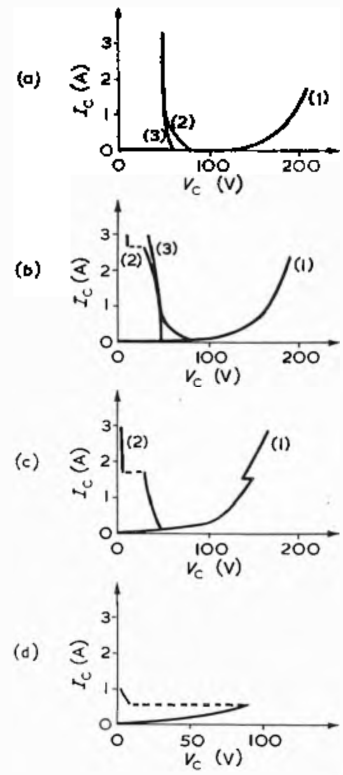


Fig. 1. Samples of a medium power junction transistor

As the form of the characteristic resembles that of semi-conductor devices[1,2] especially designed for two-state operation it may be speculated that the mechanism of the effect might be explained along similar lines. There is also the possibility that the effect is caused thermally if it is assumed that the collector current is constricted into narrow channels. In this connexion it has been noted that there exists a thermally stable negative-resistance[3] region of the reverse characteristic of a pn junction.

Yours faithfully,
B. J. Holmes,
Electronics Research Laboratory,
The British Tabulating Machine Co. Ltd.

REFERENCES

1. Moll, J. L., *et al.* PNPN Transistor Switches. *Proc. Inst. Radio Engrs.* *44*, 1174 (1956).
2. Munch, Von W. A Transistor with Thyratron Characteristics and Related Devices. *J. Brit. Instn. Radio Engrs*, *18*, 645 (1958).
3. Matz, A. W. Thermal Turnover in PN Junctions. *Proc. Instn. Elect. Engrs.* *104B*, 555 (1957).

BOOK REVIEWS

Radio Communication

By W. F. Lovering. 536 pp. 80 figs. Demy 8vo. Longmans Green & Co., New York, London. 1958. Price 60s.

THIS is a book for engineering students written from an engineer's viewpoint. Although essential mathematics are included, sometimes in Appendices at the end of a chapter, it is pleasing to find explanations and descriptions in physical and practical terms rather than a complete reliance on mathematical formulae. This is well exemplified by the chapter on Transmission Lines and feeders, where a physical picture of the conditions on the line is clearly indicated. Experience shows that many students require just this kind of explanation to supplement the mathematical treatment.

After an intial chapter introducing or revising some fundamental principles, the next four chapters cover the basic circuits which are essential in Radio Communication.

Then follows a chapter on valves and one each on radio-frequency and audio-frequency amplification. Wideband amplification is touched on, but as the title of the book is Radio Communication it is reasonable not to expect extensive coverage of subjects which belong more to television and radar. The requirements of these latter are briefly mentioned in the chapter on radio transmission and reception.

Feedback, oscillators, detectors and modulation each have a chapter and are followed by two chapters on transmitting and receiving circuits in which some representative practical circuits are given. The remainder of the book comprises a chapter each on power supplies, transmission lines (touching on waveguides), radiation and aerials, propagation, cathode-ray tubes and on transistors. Any of these could be the subject for a whole book, but the student will find them all an excellent introduction and packed with information.

The book is completed by a set of questions, with answers, relating to each chapter, and a useful index.

In a book of this size the author has to compromise between scope and completeness of detail. With so wide a scope the space devoted to each subject must be severely limited, and in some places additional diagrams would have helped.

In many instances the author points out the basic similarities and relations between what at first might appear to be quite separate problems. For example, modulation, detection and frequency changing are shown to be closely related.

He has also indicated how a problem may be solved by more than one approach, in particular, in the chapter on oscillators the action of the circuit is considered in three different ways. The book will help the student to learn that there is often more than one way of tackling a problem, and that it is important to exercise judgment in this respect, and also that in engineering many apparently diverse problems are closely related.

R. A. HERBERT

Handbook of Automation, Computation and Control. Volume 1: Control Fundamentals

By E. M. Grabbe, S. Ramo, and D. E. Wooldridge. 990 pp. 200 figs. Royal 8vo. John Wiley & Sons, Inc., New York, Chapman & Hall Ltd, London. 1958. Price 136s.

ALTHOUGH what is meant is more or less understood, no precise definition has so far been given of automation, computation or control. Each of these fields, according to viewpoint, training or particular enthusiasm, represents an approach to the evolving philosophy of Systems Engineering; a concept which is equally ill-defined as yet. The editors have sensibly chosen a title which will be interpreted widely enough to engage the attention of the most diverse audience. This is important since one of the greatest virtues of this newly recognized philosophy is its profound unifying influence, embracing servomechanisms, process control, economics, and many other previously unrelated mechanistic and sociological phenomena. This unification will only be possible if its influence reaches all who can benefit from it.

Being a handbook each chapter is a self-contained article, expertly introducing its subject, and supported by an adequate if not universally agreed choice of references. However the opportunity has been missed to include an introductory chapter showing why the selected topics are contained between the covers of this one book. Neither the foreword nor the preface, in only reiterating trite generalities, rise to the occasion.

A further, though perhaps less important editorial fault arises from the title of this volume, namely Control Fundamentals. The first sixteen chapters would better have been separated into an extremely valuable mathematical source book in its own right, leaving the remaining ten chapters aptly under the present title.

The book opens with a chapter on the theory of Sets and Relations and sharply puts the reader into the frame of mind that this is not just a handbook for ready reference, but also a text worthy of careful and patient study in its own right. The next twelve chapters cover Algebraic Equations, Matrix Theory, Finite Difference Equations, Differential Equations, Integral Equations, Complex Variables, Operational Mathematics, Laplace Transforms, Conformal Mapping, Boolean Algebra, Probability and Statistics. Each one is a skilful distillate of its subject.

Chapter 14 concisely describes most of the relevant methods of Numerical Analysis required in modern computation, while chapter 15, the longest in the book, deals with Operational Research, and includes the well-known techniques of Linear Programming and the Theory of Games, among others. The communication of intelligence is provided for in chapters 16, 17 and 18, which discuss, respectively, Information Theory, Smoothing and Filtering, and Data Transmission. It is surprising, however, that no chapter on network theory, in its own right, has been thought a subject worthy of inclusion in a handbook of this sort.

Feedback control techniques form the second half of the book starting with chapter 19 on the Methodology of Feedback Control. This chapter is indeed useful in restating the I.R.E. standard terms and symbols, and a useful table of nineteen performance specifications is given. This particular author, however, even in his title shows a flair for inventing words and their uses, and produces *horrendous*—and who knows, perhaps it means just that. Chapters 20, 21, 22 and 23 are headed Fundamentals of System Analysis, Stability, Relation between Transient and Frequency Response and Feedback Compensation respectively. These chapters are largely of selected material from the best of the literature on linear feedback control theory. All the familiar charts, response curves, tables of transfer functions, tables of transforms, root loci, etc., are included, and taken together form a very fine introductory textbook and reference. Noise, Random Inputs and Extraneous Signals is the title of chapter 24 and is a useful introduction to correlation techniques, although the subject of optimizing for minimum mean square error, or other figures of merit, is perhaps wisely only touched upon. The two final chapters deal with Non-linear Systems, and Sampled-Data Systems and Periodic Controllers. The first includes the salient features of describing-function theory and phase-plane techniques, and a little on non-linear system compensation. The chapter is a modest but valuable introduction, and also contains some very useful data on non-linear effects. Chapter 26 closes Volume I with a concise description of sampling system theory, featuring the z-transform.

Final judgment cannot be passed on this Handbook in its entirety, but if Volumes II and III maintain the standard of conciseness, and yet comprehensiveness, of Volume I, the whole should be an essential on every control, automation, or computing, engineer's bookshelf. The editors should be complimented on producing a work of remarkable evenness in style and quality from the efforts of no less than twenty-nine authors. The talent of American editors to project American talent in excellent books is truly remarkable, and is, in itself, a subject worth serious study and reflection.

K. C. GARNER

Audio Measurements

By Norman H. Crowhurst. 223 pp. 70 figs. Demy 8vo. Gernsback Publications, New York. 1959. Price $2.90.

THE author covers these measurements authoritatively and completely. Measurement techniques are described and evaluated. Test equipment, basic measurements and amplifiers are treated in detail. Measurements of output transformers, pre-amplifiers, pick-ups and arms, turntables and changers, tape recorders and microphones are also described.

Selected Papers on Quantum Electrodynamics

Edited by J. Schwinger. 424 pp. Demy 8vo. Dover Publications Inc. 1958. Price $2.45.

IN this volume the history of quantum electrodynamics is unfolded through the original words of its creators. Papers are included by Bethe, Bloch, Dirac, Dyson, Fermi, Feynman, Heisenberg, Kusch, Lamb, Oppenheimer, and many others. There are a total of thirty-four papers, twenty-nine of which are in English, one in French, three in German, and one in Italian.

International Radio Tube Encyclopaedia

600 pp. Royal 8vo. 3rd Edition. Bernards (Publishers) Ltd. 1958. Price 63s.

SINCE 1954 research in the radio valve industry has gone on unceasingly throughout the world and some further 9 000 types of radio valves have been produced, so that this present edition contains data on more than 27 500 valve types.

The scope of this encyclopaedia includes receiving valves including diodes, triodes, tetrodes, pentodes, heptodes, hexodes, tuning indicators, regulators, thyratrons, rectifiers, sub-miniature valves, television cathode-ray tubes, industrial and military type transmitting triodes, tetrodes, pentodes, cathode-ray tubes, klystrons, magnetrons, etc.

A new feature of this edition are the completely cross indexed equivalents tables, covering every receiving vale produced throughout the world with equivalents. These tables will enable users of the encyclopaedia to substitute alternatives when required.

An Introduction to Fourier Methods and the Laplace Transformation

By Philip Franklin. 289 pp. Demy 8vo. 2nd Edition. Dover Publications Inc., New York. 1958. Price $1.75.

THE author of this book is Professor of Mathematics at the Massachusetts Institute of Technology. It is an introduction to Fourier series and Laplace transforms. Applications to physical problems involving ordinary and partial differential equations are included. The reader is assumed to have a working knowledge of elementary calculus, but where topics of advanced calculus are needed, they are developed from the beginning.

There are thirty-one sets of practice problems, one for each major topic of the book. Answers to all problems are given at the end of the book.

Voice Across the Sea

By Arthur C. Clarke. 213 pp. Demy 8vo. Frederick Muller Ltd. London. 1958. 18s.

THIS book—the story of the submarine cable from its earliest days just over a century ago to the triumphant achievement of a transatlantic telephone cable in 1956 is one of the most delightful essays on a technical subject which it has ever been the reviewer's privilege to read.

Few scientists can write good English without being pompous or dull; perhaps an even smaller proportion of journalists can write accurately and with proper balance on scientific matters. Mr. Clarke has achieved both, for not only has he a deep appreciation of the romance in technical endeavour but he has the rare gift for being able to tell a technical story in a manner which is accurate, entertaining and fascinatingly instructive

This is not a history of submarine communications. The author's object, as he says, has been 'to entertain as much as to instruct' and, in doing so, the author wandered down some odd by-ways whenever the scenery has intrigued him. It will contribute little to one's understanding of telegraphy to know how Oliver Heaviside made tea, why Lord Kelvin's monocle revolutionized electrical measurements, what a Kentucky colonel was doing in Whitehall, how Western Union lost $3,000,000 in Alaska and what unlikely articles the Victorians made of gutta-percha. Yet it is precisely such trivia that make history three-dimensional and the author makes no apology for including them.

In little more than 200 pages, this is a brilliantly told story of the tragedies, romances and achievements associated with a century of submarine cables, of the personalities and the incidents which have contributed to both the triumphs and the disasters. It is the most absorbing book the reviewer has read for a very long time and it should certainly be read by everyone connected, however remotely, with electrical communication.

G. R. GARRATT

Optical Properties of Semiconductors

By T. S. Moss. 275 pp. 60 figs. Demy 8vo. Academic Press, New York. Butterworths Scientific Publications, London. 1959. Price 50s.

THIS book is primarily of interest to research physicists of academic or industrial laboratories, but will also be suitable for study by honours students concerned with solid state physics.

The opening chapters develop the theory of the optical properties of solids, including absorption by valence band electrons, free carriers and impurities, dispersion and magneto-optical effects. The book stresses the interrelation of the optical constants for any material, i.e. between the real and imaginary parts of the dielectric constants, the amplitude and phase of reflectivity and the refractive index and absorption constants. Reviews of the properties of the most interesting semiconductors are given and over 600 references are provided.

The Fundamental Principles of Quantum Mechanics with Elementary Applications

By E. C. Kemble. 611 pp. Demy 8vo. 2nd Edition. Dover Publications Inc., New York. 1958. Price $2.95.

THIS book is written by the Professor of Physics at Harvard University. It is an inductive presentation of quantum mechanics designed for the graduate student or the specialist in some other branch of physics. While some knowledge of advanced mathematics and the data of the physical sciences is desirable, treatment is simple, clear, and easily followed. Applications of the theory have been interwoven with the development of the basic mathematical structure.

Vector Analysis with an Introduction to Tensor Analysis

By A. P. Wills. 285 pp. 44 figs. Demy 8vo. 2nd Edition. Dover Publications Inc. 1958. Price $1.75.

ASSUMING no prior training in vector or tensor methods, and only a good working knowledge of calculus, the author develops the elements of vector algebra and calculus, the theory of scalar and vector fields and potential functions, transformations of volume and surface integrals, linear vector functions, coordinate systems, transformation theory, non-euclidean manifolds, and tensor theory. Numerous problems illustrate and extend each concept and enable the student to develop facility in handling vector and tensor notation.

THE ELECTRONIC COMPONENT EXHIBITION

A description, compiled from information supplied by the manufacturers, of selected exhibits to be shown at the Exhibition of the Radio and Electronic Component Manufacturers' Federation in Grosvenor House and Park Lane House, London, from 6 to 9 April. (*Figures in parentheses refer to stand numbers*)

(Voir page 243 pour la traduction français: Deutsche Ubersetzung auf Seite 249)

AVO LTD (73)

Avocet House, 92-96 Vauxhall Bridge Road, London, S.W.1.

COIL WINDING MACHINE

A new coil winding machine is being exhibited which is capable of winding almost perfect layer wound coils without the use of any medium of interleaving. The normal mechanical coupling between the headstock and the traverse of the machine has been dispensed with and both are driven by independent motors, whose relative speeds are electronically controlled by a sensing head which monitors the wire-feed angle so as to ensure that as far as practicable each turn is laid precisely beside the preceding turn. The greatly improved space factor of coils produced by this machine ensures maximum electromagnetic efficiency—a factor of particular importance for such components as relays and solenoids, especially the miniature types, while the far more precise rectilinear configuration of coils thus wound, opens up many possibilities of the use of self-supporting coils where formers were previously necessary.

EE 8751 *for further details*

BRAYHEAD (ASCOT) LTD (167, 169)

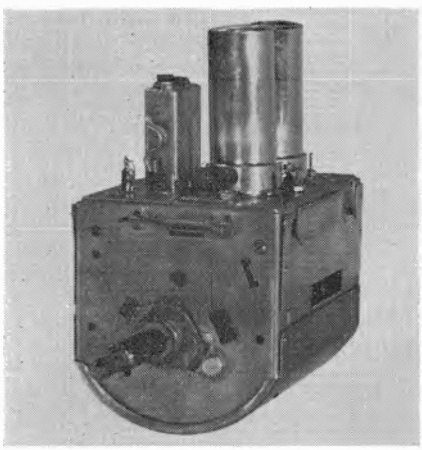
Full View Works, Kennel Ride, Ascot, Berks.

TELEVISION TURRET TUNERS

(*Illustrated below*)

The PB.1 tuner provides a high performance 10-position tuner, which is compact in construction with facilities for single and multi-channel loading, switched fine tuner range and provision for single or three-position f.m. sound. It offers a high standard of performance in respect of gain, drift, noise, stability and response shape.

The PB.3 and PB.4 tuners are the result of the combined efforts of Plessey and Brayhead. They have the same compact overall dimensions as the PB.1, providing 14 channels in the PB.3 and 18 channels in the PB.4, both retaining an additional position for the addition of a u.h.f. adaptor.



Television-f.m. switching can be provided either by a wafer switch mounted on the rear spindle extension or by a cam-operated slide switch, the latter being available in versions to suit single or multi-channel f.m. tuning.

The user has the option of many additional features giving flexibility of application; among these are:—

Printed i.f. filter, single or multi-channel f.m. tuning, built-in isolation components, narrow or wide-band fine tuner, adjustable aerial coils, inductive-coupled input circuit, double-tuned i.f. output circuit, single or dual channel loading, dual-frequency i.f. output circuit, built-in sensitivity switching, built-in aerial switching.

Provision is also made for the use of the PCC.89 valve, or the 30L15/30C15 combination.

In addition to normal U.K. channelling these tuners can be provided with channels for F.C.C. or C.C.I.R. systems, or for other special channels obtaining in overseas countries.

EE 8752 *for further details*

BRITISH PHYSICAL LABORATORIES (88)

Houseboat Works, Radlett, Hertfordshire.

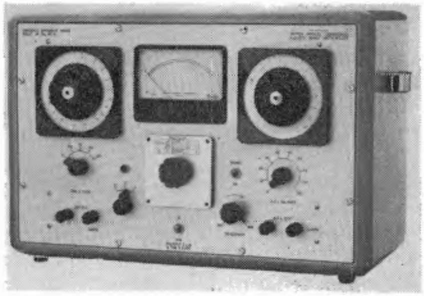
UNIVERSAL IMPEDANCE BRIDGE

(*Illustrated above right*)

This universal impedance bridge is a general-purpose instrument, covering a wide range of measurements for values of resistance, inductance and capacitance.

When used for resistance measurement, the instrument is connected as a Wheatstone bridge with an internal d.c. bridge voltage supply. In order to prevent self-heating, in particular of low wattage carbon resistors, the bridge voltage has been reduced to such a value that, at any measurement, the load of the resistor under test does not exceed 16mW. The out-of-balance voltage is amplified by means of a chopper amplifier and the indication is by means of a 4in meter, fitted with automatic overload protection.

Owing to the combination of a con-



tinuously variable resistance, and a decade resistance, an exceptionally long scale-length, namely, 60ft, is achieved. This arrangement is employed for resistance, capacitance and inductance measurement, and enables a higher reading accuracy.

As a capacitance bridge, the instrument is arranged as a modified De Saute's bridge, supplied with a 1kc/s bridge voltage from an internal oscillator. The dielectric loss is measured with a second variable control in combination with a 5-way switch. Provision is made for connecting an external d.c. source for applying a superimposed d.c. voltage to the capacitor under test. Null detection is by means of a tuned amplifier-detector, followed by a 4in indicator, fitted with automatic overload protection.

For the measurement of inductance, the instrument is converted into a Hay's bridge, the reactive standard consisting of a silvered mica capacitor; the same controls are used as for capacitance and power factor balance. By connecting an external d.c. source of supply, a current can be passed through the coil under test.

EE 8753 *for further details*

BURNDEPT LTD (121)

Erith, Kent.

CARTON CONTENTS MONITOR

(*Illustrated above right*)

The Burndept carton contents monitor type BE.251 employs a proximity switch which can sense at high speed metallic or non-metallic contents, provided the boxed article presents a sufficiently large change of electrical capacitance to the sensing probe.

The carton contents monitor kit comprises:

(a) One carton contents monitor BE.251
(b) One carton contents monitor ejector head BE.257
(c) Two carton contents monitor connectors BE.247
(d) One carton sensing head BE.258
(e) One contents sensing head BE.243

The items (d) (e) and sometimes (b) have to be designed specially to suit

HANOVER FAIR

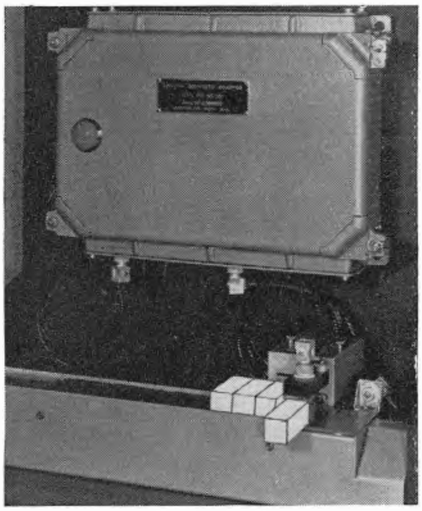
'Electronic Engineering' will occupy Stand 2 in Hall 13 at the Hanover Fair from 26 April to 13 May and visitors will be welcome.

each application depending on the size of cartons employed.

The ejector head and sensing probes are usually installed near the conveyor belt while the other units are mounted inside a cast instrument case which can be fixed to a nearby wall.

Basically, two proximity switches are used as detectors while the control unit co-ordinates their output and the action of the ejector head. One proximity switch senses (via its sensing probe) whether a box is full or empty while the other determines whether there are any boxes present on the conveyor belt. This latter refinement has been incorporated to prevent the ejector head operating unnecessarily when there are no boxes at all on the line, say, due to spacing or a production failure.

To give complete coverage for all production systems it is necessary to position the two sensing probes above each other. Where belt feed systems are used the lower probe (carton sensing probe) must be inserted within the flow line such that the articles are pressure fed over its surface.



The monitor BE.251 is capable of detecting a wide range of products in solid, powder, paste or liquid form. Foodstuffs, drugs, detergent powders, metallic pellets, etc., can be all checked before or after the carton has been sealed. Sometimes certain articles cannot possibly be detected by other methods; for example, cartons containing films could not be examined by a nucleonic method as 'fogging' of film would occur and only the capacitive method can be used in this case to determine whether a carton is full or empty. In certain cases the BE.251 can also be used for determining whether the cartons are correctly filled up to a predetermined level.

EE 8754 *for further details*

CATHODEON CRYSTALS LTD (7)

Linton, Cambridgeshire.

MINIATURE CRYSTAL OVEN

(Illustrated above right)

Specially designed for accurate temperature control of 2M type crystals, the Cathodeon miniature crystal oven is particularly suitable for close channel

mobile radio-telephone systems where precision frequency control is essential but space is at a premium. Frequency stability better than ±·0005 per cent over a wide temperature range is readily achieved. Compact design, low power consumption, and rapid heating from cold, are some of the special features.

The oven operates from a 6V or 12V supply with a power consumption of 4·6W and standard temperature settings of 75°C, 80°C and 85°C are available. The heating time is less than 5min from +20°C to +85°C. The weight is 1·5oz.

EE 8755 *for further details*

SUB-MINIATURE CRYSTALS

A new range of plug-in subminiature crystal units has been introduced. This unit, type 2MM, is especially suitable for use with transistors and other miniature components. They are available in a frequency range from 5 to 75Mc/s (fundamental 5 to 20Mc/s and series resonant overtone 15 to 75Mc/s) with a performance of ±·005 per cent from −55°C to 105°C

EE 8756 *for further details*

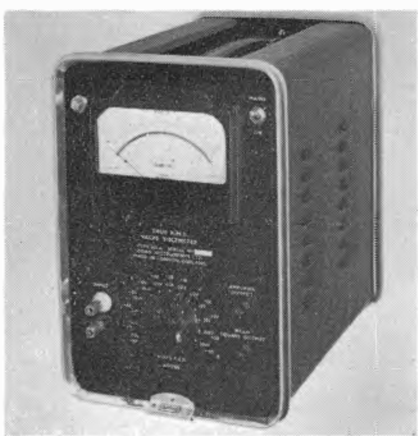
DAWE INSTRUMENTS LTD (66)

99 Uxbridge Road, Ealing, London, W.5.

TRUE R.M.S. VALVE-VOLTMETER

(Illustrated below)

Unlike the usual valve-voltmeters which are average or peak reading instruments calibrated to read r.m.s. values, the type 612 true r.m.s. valve-voltmeter measures accurately the effec-



tive value of waveforms other than sinusoidal. Accurate r.m.s. reading can be obtained for voltages or currents associated with non-linear devices or circuits. Thirteen voltage ranges are provided with full scale meter deflexions of 300μV, 1mV, 3mV, 100mV, 300mV, 1V, 3V, 10V, 30V, 100V and 300V. The amplifier response does not vary more than ±3 per cent from 15c/s to 150kc/s and ±5 per cent 5c/s to 500kc/s.

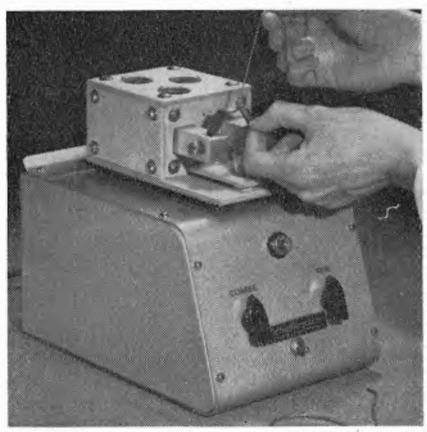
EE 8757 *for further details*

HIGH OUTPUT A.F. OSCILLATOR

(Illustrated above)

This oscillator is of the resistance capacitance tuned type covering the frequency range 20c/s to 20kc/s. A feature of the oscillator is the high output (6W) matching into a load impedance of 15Ω or 600Ω, i.e. ample power to drive loudspeakers or small vibration generators direct. Accuracy is ±1 per cent on the 200c/s to 2kc/s range with ±2 per cent accuracy on other ranges.

EE 8758 *for further details*



ELECTROTHERMAL ENGINEERING LTD (135)

270 Neville Road, London, E.7.

PRECISION BRAZING EQUIPMENT

(Illustrated above)

This equipment has been developed to provide an easily operated, portable and compact unit for use with silver brazing alloys. It is of especial use for the hard soldering of resistance wires, such as nickel chrome and related alloys, where soft solder is either difficult to use or has too low a melting point for the work in hand, and where spot welding gives too weak a joint.

Although mainly designed for fine work (from ·001in diameter) the unit can be used to join wire as heavy as 10s.w.g. (0·128in diameter).

The consumption of the unit depends upon the control setting; the maximum is approximately 500W. Solder is heated by means of an expendable element which is easily replaced. Receptacles for flux and solutions for flux removal are conveniently placed for the operator, the solutions being heated. The equipment is suitable for use on any a.c. supply from 200 to 250V without adjustment, and other a.c. supplies to special order.

EE 8759 *for further details*

FORTIPHONE LTD (152)

92 Middlesex Street, London, E.1.

SUB-MINIATURE POTENTIOMETERS

The type VS32 sub-miniature potentiometer and s.p. switch is of extremely small dimensions and is intended for use in headborne hearing aids, etc. It is designed for frontal or edgewise mounting and is secured by a central screw and nut and has an overall diameter of 0·312in and depth of 0·15in. It can be supplied in resistance values between 1kΩ and 3MΩ with a power rating of 50mW.

The type PSR preset potentiometer is available for conventional wired or printed circuit applications. The overall diameter of 0·45in and the case depth 0·08in. The resistance range and power rating are the same as for the type VS32.

The type PSS is a preset sliding potentiometer of simple design and construction having an overall length of 1·1in and an overall width of 0·28in. The resistance range is 1kΩ to 3MΩ and the power rating 50mW.

EE 8760 *for further details*

GEO. L. SCOTT & CO. LTD (133)

Cromwell Road, Ellesmere Port, Cheshire.

LAMINATIONS AND CORES

A range of sub-miniature transformer laminations is now available covering overall dimensions from 1in by ¾in down to ⅜in by ⅜in, for use in transistor circuits and where space is at a premium.

Toroidal cores, using nickel iron materials made by the powder metallurgy process, are available on high temperature ceramic bobbins, to RCL 193 sizes, on which windings can be directly applied. The consistency of characteristics and the high performance to be gained by using these alloys is apparent where, for instance, matched pairs are needed for magnetic amplifiers. Silicon iron toroidal cores in materials of ·013in, ·004in and ·002in thickness are also available to RCL 193 as well as to customers' standards, for low and high frequency power sources.

Fast switching and high permeability tapes down to ·0003in thick are also produced on ceramic bobbins for switch matrices and other high performance magnetic devices.

EE 8761 *for further details*

HINCHLEY ENGINEERING CO. LTD (91)

Pans Lane, Devizes, Wiltshire

ISOLATING TRANSFORMERS

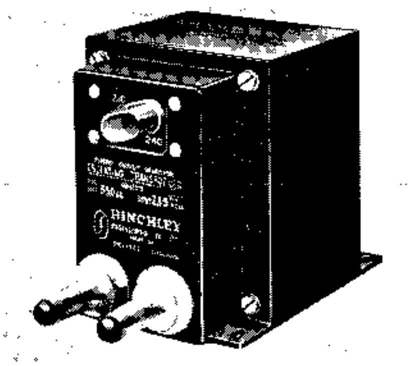
(Illustrated below)

In addition to a large range of transformers for radio and television applications, etc., a range of standard isolating and auto-transformers for laboratory, workshop and service department use will also be shown. These transformers are all provided with rapid-change fused output voltage selectors, and are optionally available with Mycalex snap-action terminals.

Of particular interest, in view of the growing number of domestic electrical appliances for use in the workshop and the garden, are two versions of the Hinchley 'Safetybox' which are portable isolating transformers enclosed in stout steel cases (either 1 : 1 ratio for isolation of existing equipment; or, preferably, 110V or 55V output with centre-point earthed).

These have been designed specifically for large-quantity production, so that the ultimate cost to the consumer will make safety a more economic necessity.

EE 8762 *for further details*



JAMES A. JOBLING & CO. LTD (158)

Wear Glass Works, Sunderland.

METAL OXIDE RESISTORS

These high precision, high stability resistors are made from 'Pyrex' brand rod to which a thin film of metallic oxide is fused. The core and film are inseparably impervious and stable under the most advance conditions and their noise level is extremely low.

The standard tolerance of these resistors is 1 per cent and they are manufactured to customers' exact requirements rather than to preferred values, in a range between 100Ω and 250kΩ, and in power ratings between 0·5W and 2W.

EE 8763 *for further details*

LABGEAR LTD (110)

Willow Place, Cambridge.

NON-DESTRUCTIVE FLASH TESTER

(Illustrated above right)

This equipment is designed for testing the insulation of circuits and components such as low value capacitors.

The standard model provides a.c. test voltages of 250V, 500V, 1kV, 1·5kV and 2kV. A feature of the instrument is that it allows a component to be tested in

excess of its normal working voltage, and if breakdown should result, the component is completely undamaged and may be re-employed at a lower potential, which allows components to be graded without the risk of destruction.

EE 8764 *for further details*

LUSTRAPHONE LTD (111)

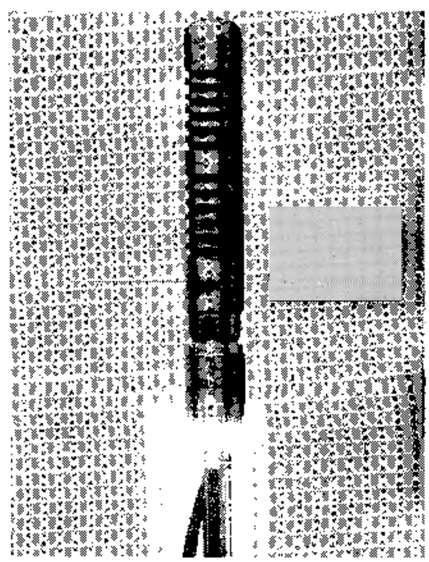
St. George's Works, Regents Park Road, London, N.W.1.

STEREOPHONIC MICROPHONE

(Illustrated below)

This microphone unit consists of two identical ribbon velocity microphones mounted vertically in line. The top half of the unit, which contains one of the microphones, is so arranged that it can be continuously rotated through an angle of approximately 100° from the 'in line' position, thus enabling the optimum setting for any contingency (of stereophonic transmission, recording, broadcasting, sound re-inforcement etc.) to be achieved. The lower half of the microphone is placed in a fixed position.

The microphone is connected to a switch unit by means of a 4 way screw-type connector. The advantage of this method is that the sensitive microphone



head can be instantly detached and stored in a safe place, without disturbing the set-up and connexions to the main equipment.

A specially designed switch is provided with three positions:

(1) Off position in the centre.

(2) Stereo connexion on the extreme of one side.

(3) The other extreme position on the switch connects the two ribbon units in series and thereby changes the microphone into a straightforward ribbon velocity microphone of high sensitivity and a figure '8' polar characteristic.

A phase reversing switch is provided at the rear of the switch unit so as to give instantaneous changeover to achieve the correct phasing on any equipment.

When the microphone is set in position as described in (3) above, and with both units 'in-line', operation of the phase reversing switch changes the microphone into a noise eliminating microphone of substantial efficiency, responding only to a sound source at close distance.

In conjunction with the alternative positions of the phase switch, intermediate relative positioning of the microphone units gives a variety of directional effects which can prove beneficial in difficult acoustic conditions.

The impedance of each microphone unit in the standard model is 20Ω and the overall impedance of the combined instrument switched to the series position is 40Ω.

EE 8765 *for further details*

McMURDO INSTRUMENT CO. LTD (32)

Victoria Works, Ashtead, Surrey.

VALVEHOLDERS AND CONNECTORS

A new range of pen nib valveholders using single blank mouldings in both B9A and B7G versions for printed circuit and conventional wiring is available. These are designed to accept simple spring valve retainers, the fixing of which is integral with the metal mounting of the valveholder. In addition the valves can be screened by cheap rolled up screening cans which are a close fit around the valve envelope and are retained and electrically earthed by the spring retainer.

The Pakonector range of plugs and sockets has been added to by the introduction of an 8 way version which is similar in size to a B7G valveholder. The same cable connexion and potting service is available on the 8 way as on the original 18 way versions.

EE 8766 *for further details*

MULLARD LTD (77, 67)

Mullard House, Torrington Place, London, W.C.1.

110° CATHODE-RAY TUBES

Both 21in and 17in sizes are being produced, under the type numbers AW53-88 and AW43-88 respectively

The new 110° tubes are substantially shorter than the equivalent sizes of tubes currently in use, and will make possible reductions in the front-to-back dimensions of television receiver cabinets.

In the case of the 21in screen, the 110° tube is over 8½in shorter than its 70° counterpart, and 5in shorter than a 90° type. The 17in 110° tube shows reductions of approximately 6½in and 3in, respectively, over the equivalent size of 70° and 90° types.

The new tubes have high-efficiency metal-backed phosphors, and provide excellent brightness and contrast under high ambient light levels. They are electrostatically focused, and incorporate the 'straight' type of electron gun, which dispenses with an ion-trap magnet.

EE 8767 *for further details*

PAINTON & CO. LTD (29)

Kingsthorpe, Northampton.

MINIATURE HIGH STABILITY RESISTORS

These extremely small high stability carbon resistors (type 70) measure only ⅜in by 1/10in (approximately) and are thus suitable for use in miniature transistorized equipments and particularly printed circuit board applications. The leads are pre-tinned to facilitate dip soldering. The resistance range covers standard values from 5Ω to 250kΩ at a tolerance of 5 per cent but tolerances of 1 per cent and 2 per cent can when necessary be provided over most of the range.

EE 8768 *for further details*

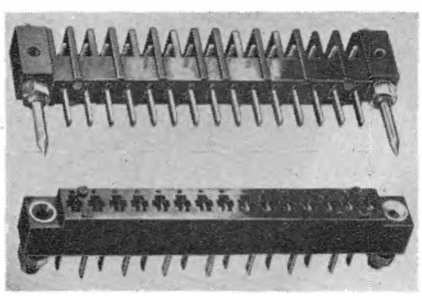
PRINTED CIRCUIT CONNECTORS

(Illustrated below)

A number of printed circuit connectors have been introduced including a stoutly constructed 15-way plug and socket fitted with steel dowels. In company with most of the Painton printed circuit connectors the contacts are on a 0·1in module.

Also in production is an edge connector available with up to 60 contacts in three groups of 20 with 0·1in centres, the contacts are of a unique and patented design and tests have shown that this form of contact permits a very long board insertion life.

EE 8769 *for further details*

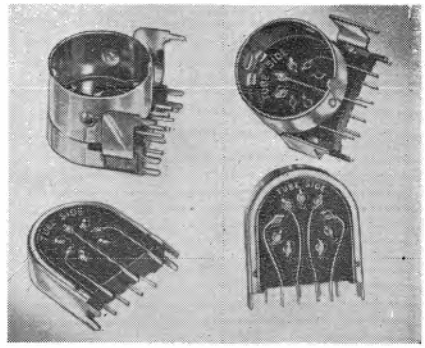


RIGHT-ANGLE VALVEHOLDERS

(Illustrated above right)

These components have been designed to permit horizontal mounting of valves, so facilitating the production of more compact printed circuit assemblies and simplifying heat sink arrangements.

The contacts are silver-plated beryl-



lium copper and are of such a length to suit the most commonly used printed board thicknesses. The correct insertion and withdrawal pressures are maintained with minimum contact resistance.

EE 8770 *for further details*

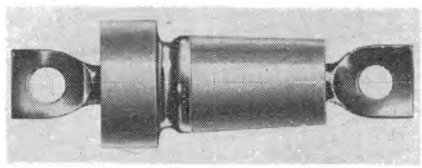
STEATITE AND PORCELAIN PRODUCTS LTD (76)

Stourport-on-Severn. Worcestershire.

TERMINAL SEALS

(Illustrated below)

Following the preliminary announcement last year, the new range of terminal seals will be shown designed to conform with the requirements of the



Services' specification PCL 331. These are made of a high alumina ceramic and the stems are firmly sealed into position by means of glaze fusion. External metallized bands are provided by a quality approved nickel metallizing process to permit of soft-soldering directly into metal cans for all hermetic sealing applications.

The standard ranges of nickel metallized seals, quality approved to the H1 category, are still available from stock and a sample making service can provide prototypes of new designs at short notice. A new emphasis has been placed on improved facilities for assembling metallized ceramic products with appropriate metal fittings to cover a wide variety of applications ranging from miniature lead-outs to large, high-voltage capacitor terminals.

EE 8771 *for further details*

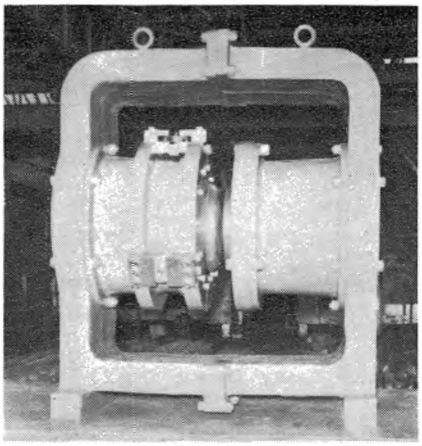
SWIFT LEVICK & SONS LTD (156)

Clarence Steel Works, Sheffield 4.

LARGE PERMANENT MAGNETS

(Illustrated overleaf)

Very large and stable permanent magnets are now being produced with fields of high homogeneity for nuclear magnetic resonance purposes. The photograph shows such a magnet designed and produced for research at Leeds University. This magnet weighs 2 tons 6 cwt, measures 4ft by 3½ft by 2ft overall, and is permanently energized by Columax alloy, which is the most stable and



powerful magnet material produced commercially anywhere in the world. The magnet has a gap field strength of 9 700 oersted over a gap volume of 1·4in long by 10in diameter, and the uniformity of field is better than 1 part in 10^6 over a central volume of about $1cm^3$. Special mechanical adjustments are provided on one of the iron pole pieces, thus enabling restricted movement of the pole face in any plane for final uniform adjustment of the gap field strength.

A one-sixth scale model of this magnet will be exhibited.

EE 8772 *for further details*

TELEPHONE MANUFACTURING CO. LTD (21)

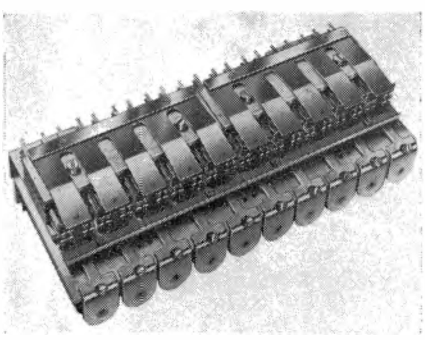
Hollingsworth Works, London, S.E.21.

MULTIPLE RELAY

(Illustrated below)

The new TMC 10-way multiple relay will be shown. In this new relay many unique and novel manufacturing methods have been employed which will bring about a low cost together with an increase in reliability. The TMC multiple relay employs a common yoke integral with mounting brackets, and is intended to replace telephone type relays in control circuits where cost is of prime importance.

Contact springs of phosphor-bronze, with silver bimetal contacts in the crossed radii technique, and automatic machine adjustment and card operation, ensure a constant contact pressure hitherto available only in the expensive precision relays. An efficient magnetic circuit ensures that magnetic interaction between relays is virtually non-existent, while the common yoke mounting



brackets greatly reduce the number of piece parts and eliminate the need for separate mounting pieces.

The price of this multiple relay is such that it will frequently be found more economical to fit a 10-way unit rather than a lesser number of conventional relays.

EE 8773 *for further details*

WHITELEY ELECTRICAL RADIO CO. LTD (81)

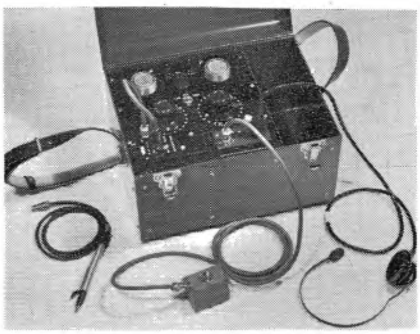
Radio Works, Victoria Street, Mansfield, Nottinghamshire.

CABLE TEST SET

(Illustrated below)

This tester locates the position of breaks or short-circuits in sheathed multiple conductor cables to within a fraction of an inch. Repairs can then be carried out with a minimum of disturbance to the cable sheath.

The principle of operation is as follows: In the case of a broken conductor



an a.f. voltage is applied between one end of the conductor and earth; the electrostatic field between the live section of the conductor and earth can then be detected by running a capacitive probe along the cable. The probe is connected to the input of a portable transistor amplifier, which feeds a headset thus giving an audible indication of the break position. All other conductors in the cable should be earthed, other than the one being tested.

An inductive probe is used in place of the capacitive probe for locating shorts between conductors. The oscillator will feed a signal to the conductor only up to the point where the short occurs and the probe will detect the electromagnetic field up to this point.

The test panel is provided with rotary switches to select individual conductors in the cable. Conductors other than the one selected are earthed. Sockets are fitted to accommodate the cable plugs. For initial determination of a faulty conductor a lamp type continuity tester is provided.

The valve oscillator is battery operated and has an output power of 10mW with an output impedance of 300Ω. Its frequency is 1kc/s ±2½ per cent. The type of oscillation can either be continuous or interrupted depending upon the setting of the potentiometer.

The transistor amplifier operates from a 4½V battery contained in the case and has an overall gain of 70dB when terminated with a 300Ω non-inductive load.

The inductive probe is tuned to 1kc/s ±2½ per cent.

EE 8774 *for further details*

THE ZENITH ELECTRIC CO. LTD (8)

Zenith Works, Villiers Road, Willesden Green, London, N.W.2.

REGULATING TRANSFORMERS

(Illustrated below)

A new 'Variac' model 50-B/2B having a dual output will be shown. This is designed for a 230V 50c/s supply providing an output variable between zero and 270V, maximum rated current 20A, which represents the total that may be taken from the two brushes but the load can, of course, be distributed to suit an application calling for example; Output No. 1—12A, Output No. 2—8A. The two brush, dual output 'Variacs' have been available in the series '100' and 'V30H' for some time and also in the smaller toroidal range in the series '200'.

These smaller types have now been re-designed using parts from the 'V5' series, and allows a more compact example to be produced.

EE 8775 *for further details*

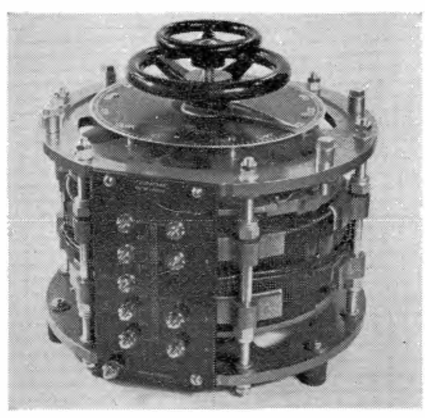
EARTH LOOP TESTING EQUIPMENT

This is housed in a metal cabinet complete with carrying handle and a pocket for an inlet cable and plug adaptor. It is designed to completely cover all tests required by I.E.E. Regulation No. 507 and to operate from a standard single-phase supply giving an output of 25A at any voltage between zero and 40V.

A Zenith double wound, step down transformer is used with a 'Variac' across the primary, a double range rectifier voltmeter being provided together with an ammeter, there is a 3-pin 13A and a 3-pin 15A outlet socket on the front of the casing for earth core tests of portable appliances. The 40V is available for most tests but 6V is included as this may be more convenient for short earth loops such as portable tools and other similar appliances.

A link is fitted for testing fixed appliances for which the 6V would be employed as the earth lead is comparatively short and should not exceed 0·1Ω representing about 3V.

EE 8776 *for further details*



Short News Items

Interlab Inc. of New York have instituted a new type of marketing and consulting service for members of the British electronic industry, the purpose of which is to help British manufacturers to explore the U.S. market for their products, and set up marketing and licensing arrangements or negotiate manufacturing licences in respect of American electronic products. This idea was tried out on an experimental basis in the spring of 1958 and was the outcome of the Company's four years of marketing experience in Britain. Interlab work closely with the Commercial Department of the British Consulate in New York as well as with the Trade Promotion Centre to give a more specialized service. Further information can be obtained by applying direct to Interlab Inc., 437 Fifth Avenue, New York 16, N.Y., U.S.A.

Marconi's Wireless Telegraph Co. Ltd have been awarded contracts from the Norwegian P.T.T. Administration and the Royal Board of Swedish Communications for a combined multi-channel radio-telephone and television link between the Norwegian capital, Oslo, and Karlstad (Sweden). A further section, carrying multi-channel radio-telephony only, is to be installed in Sweden between Karlstad and Arvika. Between Oslo and Karlstad two two-way paths and one reversible one-way path are to be constructed. Both will operate in the s.h.f. band (4 000Mc/s). One of the two-way paths will carry 600 channels of radio-telephony while the reversible single-path system will carry television signals. The second two-way path is to function as a standby for either. Broadcast speech circuits and control and supervisory signals will be transmitted via an additional radio link operating in the 400Mc/s band. The path carrying the television signals will enable programmes emanating from Oslo to be fed into the Swedish link and so into the Eurovision network; by reversing the path, Oslo can take programmes from Sweden or the Eurovision network. On the Norwegian side of the border, the twin two-way and the single reversible terminal equipments will be sited at Oslo, with the first repeater at Tryvasshogda and a second at Runddellen. In Sweden, one repeater will be sited at Blabarskullen with a terminal at Karlstad. The section from Karlstad to Arvika (multi-channel radio-telephony only will also be a twin-path system with one path carrying 600 telephone channels and the second as a standby. The distances involved are: Oslo to Karlstad, 190km, and Karlstad to Arvika, 40km. A survey of the sites has been carried out and the system is expected to be in operation by May 1960.

The Radio Industry Council announces the award of six premiums of 25 guineas each for articles on radio and electronics published in the technical press during 1958. The premiums are awarded as follows:

'A New High-Frequency High-Power Amplifier' by V. J. Tyler, B.Sc., A.M.I.E.E. (Marconi's Wireless Telegraph Co. Ltd). Published in *The Marconi Review*, 3rd Quarter, 1958.

'New Types of D.C. Amplifiers'—Parts 1 and 2—by D. J. R. Martin, B.Sc., A.Inst.P. (The Mining Research Establishment). Published in *Electronic and Radio Engineer*, Jan./Feb. 1958.

'A Survey of Microwave Radio Communication' by W. J. Bray, M.Sc. (Eng.), M.I.E.E. (Post Office Engineering Department). Published in *Electronic Engineering*, May 1958.

'All Travelling-Wave Tube Systems' by S. Fedida, B.Sc.(Eng.)(Hons.), A.C.G.I., A.M.I.E.E. (Marconi's Wireless Telegraph Co. Ltd). Published in *Electronic Engineering*, May 1958.

'Electronic Developments at Very Low Temperatures' by E. Mendoza, M.A., Ph.D. (Manchester University). Published in *British Communications and Electronics*, April 1958.

'Analogue Computation' by E. Lloyd Thomas, B.Sc., A.C.G.I., M.I.E.E., A.F.R.Ae.S. (R. B. Pullin & Co. Ltd). Published in *British Communications and Electronics*, May 1958.

The number of articles submitted for consideration was fifty-nine. The presentation of awards will be made at a luncheon on Wednesday, 15 April, by the new director of the R.I.C., Air Marshal Sir Raymond G. Hart, with Mr. W. M. York, chairman of the public relations committee, presiding.

Ampex Electronics Ltd, a new subsidiary of the Ampex Corporation, California, will start the manufacture in Britain of magnetic tape recording equipment. The factory will be in Reading with initial production scheduled for early summer of instrumentation magnetic tape equipment having applications in many industrial and research fields. Heading the Company, a wholly owned subsidiary, will be Dr. Peter Axon, formerly of the research department of the BBC, who is an authority on audio and video magnetic recording research and equipment.

Gilmoor Control Systems Limited of London, a Company owned by Electroflo Meters Co. Ltd, is engaged in the manufacture of a remote control and indication system employing a v.h.f. radio link. The system is employed for remote control and the position indication of sixteen water pumps raising and transferring water from an oasis to an installation of the Qatar Petroleum Company in the Persian Gulf. The control signalling takes place at 173·5Mc/s while indications back to the control station are transmitted at 160·3Mc/s with two-way speech communication on a frequency modulated channel at 167Mc/s. The control station and the controlled pumps are seven miles apart.

The Electronic Engineering Association held their annual general meeting in London on 23 February and elected Mr. L. T. Hinton (Standard Telephones & Cables Ltd) chairman in succession to Mr. F. S. Mockford (Marconi's Wireless Telegraph Co. Ltd). Mr. R. R. C. Rankin (Mullard Ltd) was elected vice-chairman. Mr. Hinton's connexion with radio communication goes back to experience in the Royal Engineers (Signals) in the 1914-18 war. He is now the manager responsible to his Company for all relations with trade associations, commercial and technical, and contractual relations with the Government through the Federation of British Industries. The Council was enlarged from twelve to fourteen members. All the 1958 members were re-elected with the exception of Kelvin & Hughes Ltd, who stood down, and newly-elected members were Automatic Telephone & Electric Co. Ltd, Siemens Edison Swan Ltd, and Ultra Electric Ltd. Members not standing for election besides Kelvin & Hughes Ltd, were G.E.C., Marconi Instruments Ltd, Marconi International Marine, Philips Electrical Ltd and Rank Cintel Ltd.

The Institution of Electrical Engineers held their annual dinner at Grosvenor House on Thursday, 26 February, 1959, with the President, Mr. S. E. Goodall, in the chair. The principal guest was the Lord Chancellor, the Rt. Hon The Viscount Kilmuir, who proposed the toast of 'The Institution', to which the president replied. The toast of 'Our Guests' was proposed by Mr. T. E. Goldup (Immediate Past-President) and the reply was given by Sir Norman Kipping (Director-General, Federation of British Industries).

Erratum. In the article 'An Introduction to Matrices' by J. S. Bell and K. Brewster in the February issue, the two lines preceding the 'Conclusion' on page 102 should read:

$$= \frac{10^3}{0{\cdot}198\angle 1^\circ}$$
$$\simeq 5{\cdot}05\text{k}\Omega$$

Meetings this Month

THE BRITISH INSTITUTION OF RADIO ENGINEERS

Date: 22 April. Time: 6.30 p.m.
Held at: London School of Hygiene and Tropical Medicine, Keppel Street, Gower Street, London, W.C.1.
Lecture: The Application of Magnetic Resonance to Solid State Electronics.
By: D. J. E. Ingram.

South Midlands Section

Date: 2 April. Time: 7.0 p.m.
Held at: Winter Gardens, Malvern.
Lecture: A Simple High-Quality F.M. Broadcast Receiver employing a Pulse-rate Discriminator.
By: P. J. Baxandall.

Scottish Section

Date: 17 April. Time: 6.30 p.m.
Held at: Department of Natural Philosophy, The University, Drummond Street, Edinburgh.
Annual General Meeting.
Lecture: Stereophonic Sound and Electrostatic Loudspeakers.
By: D. T. N. Williamson.

Merseyside Section

Date: 22 April. Time: 7 p.m.
Held at: University Club, Liverpool.
Annual General Meeting.

North Eastern Section

Date: 8 April. Time: 6 p.m.
Held at: Institution of Mining and Mechanical Engineers, Neville Hall, Westgate Road, Newcastle upon Tyne.
Lecture: Radio Exploration of the Galaxy.
By: J. Baldwin.

North Western Section

Date: 2 April. Time: 6.30 p.m.
Held at: Reynolds Hall, College of Technology, Sackville Street, Manchester 1.
Lecture: Principles of Transistor Circuitry.
By: B. R. A. Bettridge.

BRITISH SOUND RECORDING ASSOCIATION

Date: 17 April. Time: 7.15 p.m.
Held at: Royal Society of Arts, John Adam Street, Adelphi, London, W.C.2.
Lecture: The Quest for Quality.
By: P. Ford.

THE INSTITUTION OF ELECTRICAL ENGINEERS

All London meetings, unless otherwise stated, will be held at the Institution, commencing at 5.30 p.m.

Ordinary Meeting

Date: 2 April.
Lecture: Women in Engineering.
Opened by: Sir Willis Jackson.
Date: 23 April.
Kelvin Lecture: The Geophysical Year 1957-58.
By: Sir David Brunt.

Measurement and Control Section

Date: 7 April.
Lecture: An Electron Trajectory Tracer for use with the Resistance Network Analogue.
By: M. E. Haine and J. Vine.
Date: 21 April.
Discussion: The Problem of Maintenance of Electronic Equipment in the Process Industries.

Radio and Telecommunication Section

Date: 8 April.
Lecture: The New Post Office 700-Type Telephone.
By: F. E. Williams and F. C. Carter.
Date: 17 April.
Lecture: Engineering Aspects of Commercial Television Programme Presentation.
By: T. C. Macnamara and B. Marsden.
Date: 27 April.
Lecture: The Field Strengths Required for the Reception of Television in Bands I, III, IV and V.
By: G. F. Swann.

Utilization Section

Date: 16 April.
Lecture: An Electromagnetic, Variable-Ratio, Torque Convertor.
By: D. O. Bishop and G. S. Brosan.

East Midland Centre

Date: 7 April. Time: 6.30 p.m.
Held at: College of Arts and Crafts, Nottingham.
Lecture: Subscriber Trunk Dialling.
By: D. A. Barron.

Cambridge Radio and Telecommunication Group

Date: 14 April. Time: 7 p.m.
Held at: Cavendish Laboratory, Free School Lane, Cambridge.
Lecture: Designing Electronic Products.
By: J. R. Brinkley.
Lecture: The Application of Counting Techniques; and Some Unusual Industrial Talks.
By: C. C. H. Washtell.
Lecture: Measurement of Mechanical Parameters using Electrical Screening.
By: B. E. Noltingk.
Lecture: Some Aspects of Pulse Generation, Recording and Analysis.
By: J. Barron.
Lecture: An Unusual Circuit Element.
By: R. Nettleship.
Lecture: Some Features of Multi-Channel Recording in Electromedicine.
By: N. Bett.

East Anglian Sub-Centre

Date: 20 April. Time: 7.30 p.m.
Held at: Assembly House, Norwich.
Annual General Meeting.
Lecture: High-Quality Sound Reproduction.
By: J. Moir.

North-Eastern Centre

Date: 13 April. Time: 6.15 p.m.
Held at: Royal Station Hotel, Newcastle upon Tyne.
Annual General Meeting and Conversazione.

North-Eastern Radio and Measurement Group

Date: 6 April. Time: 6.15 p.m.
Held at: Rutherford College of Technology, Newcastle upon Tyne.
Annual General Meeting and Film Evening.

North Midland Centre

Date: 7 April. Time: 7 p.m.
Held at: Royal Station Hotel, York.
Lecture: The Relation between Picture Size, Viewing Distance and Picture Quality.
By: L. C. Jesty.

Sheffield Sub-Centre

Date: 15 April. Time: 6.30 p.m.
Held at: Grand Hotel, Sheffield.
Lecture: Subscriber Trunk Dialling.
By: D. A. Barron.

North-Western Centre

Date: 7 April. Time: 6.15 p.m.
Held at: Engineers' Club, 17 Albert Square, Manchester.
Lecture: H.V. D.C. Transmission, with special reference to the Cross Channel Cable.
By: F. J. Lane.

North Ireland Centre

Date: 14 April. Time: 6.30 p.m.
Held at: David Keir Building, Queen's University, Stranmillis Road, Belfast.
Lecture: The Use of Analogue Computing Elements in the Design of Automatic Control Systems.
By: Professor J. C. West.

South-West Scotland Sub-Centre

Date: 22 April. Time: 7 p.m.
Held at: The Institution of Engineers and Shipbuilders, 39 Elmbank Crescent, Glasgow, C.2.
Lecture: A Train Performance Computer.
By: E. Bradshaw, M. Wagstaff, F. Cooke.

South Midland Radio and Measurement Group

Date: 27 April. Time: 6 p.m.
Held at: James Watt Memorial Institute, Birmingham.
Lecture: Stereophonic sound.
By: K. N. Hawke.

North Staffordshire Sub-Centre

Date: 16 April. Time: 7 p.m.
Held at: Duncan Hall, Stone.
Lecture: Subscriber Trunk Dialling.
By: D. A. Barron.

Southern Centre

Date: 8 April. Time: 6.30 p.m.
Held at: S.E.B. Canteen, Drayton, Portsmouth.
Annual General Meeting.
Lecture: Rockets and Satellites.
By: R. L. F. Boyd.

Western Centre

Date: 13 April. Time: 6 p.m.
Held at: St. Mary's College, Cheltenham.
Lecture: Transistors in Communication and Control Equipment—A General Survey.
By: E. Wolfendale.

THE TELEVISION SOCIETY

Date: 23 April. Time: 7 p.m.
Held at: Cinematograph Exhibitors' Association, 164 Shaftesbury Avenue, London, W.C.2.
Lecture: Design of Experimental Tuners for Bands 4 and 5 Television Receivers.
By: K. H. Smith.

THE SOCIETY OF INSTRUMENT TECHNOLOGY

Date: 28 April. Time: 6 p.m.
Held at: Manson House, Portland Place, London, W.1.
Lecture: Measurements in the Photographic Industry.
By: M. E. F. Howarth and P. B. Watt.

PUBLICATIONS RECEIVED

THE NICKEL BULLETIN, JANUARY 1959, contains abstracts which give an apt illustration of the immense importance of nickel in every branch of industry. In the general field of nickel plating, reference is made, for example, to electrodeposition of black nickel, tin-nickel and hard carbide/nickel coatings. The section dealing with non-ferrous alloys includes abstracts of papers on physical and magnetic properties and the influence of nickel content of copper alloys on susceptibility to intergranular cavitation. The issue also contains the first of the quarterly reviews of recent patent literature. The Mond Nickel Co. Ltd, Thames House, Millbank, London, S.W.1.

DIGITIZERS AND OTHER EQUIPMENT FOR DATA PROCESSING is a publication by Hilger and Watts dealing mainly with digitizers which are for use when measurements of mechanical movements which may be the analogues of variables, are wanted in digital form. Hilger and Watts Ltd., 98 St. Pancras Way, Camden Road, London, N.W.1.

ABRIDGED VALVE DATA is a quick reference brochure, which is issued annually and gives abridged data for all English Electric Valves up to the date of publication. Technical Publications Department, English Electric Valve Co. Ltd. Chelmsford, Essex.

FILMS FOR INDUSTRY (1959-1960) EDITION supersedes all earlier editions of this catalogue and includes a large number of newly acquired films from various sources. The 612 films listed in this catalogue have been selected under the guidance of an Advisory Panel whose members include representatives from industry and Government Departments concerned with industry. Central Office of Information, Hercules Road, Westminster Bridge Road, London, S.E.1. Price 2s. 6d.

PERMANENT MAGNETS SUMMARIZED by F. G. Tyack, A.M.I.E.E., is a booklet recently prepared by James Neill & Company (Sheffield) Limited, manufacturers of 'Eclipse' Permanent Magnets, Magnetic Chucks and Magnetic Tools, and deals with the application of permanent magnets. This publication contains technical information on magnet design and materials and its contents are in convenient sequence for reference, while the illustrations and technical diagrams are captioned for easy reading and reference.

AN APPLICATION REPORT ON D.C. POWER SUPPLY CIRCUITS USING SILICON RECTIFIERS has been published by Texas Instruments Limited. This is the fourth in the series and draws attention to the particular design features of circuits using silicon rectifiers. The more common single-phase and three-phase rectifier circuits are described. Design procedures are discussed for circuits using capacitor-input and choke-input filters and for additional filter sections. For convenience, graphs used in the design procedures are collated at the end of the report. Texas Instruments Limited. Dallas Road, Bedford.

CHILTON STYLE 'C' ROTARY SWITCHES catalogues a very large number of different switches, and a great many accessories in this third edition by Chilton Electric Products Ltd. Hungerford, Berks.

(Pages 243 to 254 French and German section)

Le Salon BRITANNIQUE DE LA PIÈCE DÉTACHÉE ÉLECTRONIQUE

Description, basée sur des renseignements fournis par les fabricants, d'un choix de pièces détachées qui seront exposées au Salon de la Fédération britannique des Fabricants de Pièces Détachées et Accessoires radio-électriques et électroniques, à Grosvenor House et à Park Lane House, Londres, du 6 au 9 avril 1959.
(Les chiffres entre parenthèses se rapportent aux numéros des stands). *(Traduction des pages 236 à 240)*

AVO LTD (73)

Avocet House, 92-96 Vauxhall Bridge Road, London S.W.1

MACHINE À BOBINER

Une nouvelle machine à bobiner sera exposée par cette société. Cette machine est capable d'enrouler des bobinages en couches superposées presque parfaites, sans l'emploi d'un moyen quelconque d'interfoliage. Le couplage mécanique normal entre la poupée fixe et la pièce transversale de la machine a été supprimé et toutes deux sont actionnées par des moteurs indépendants dont les vitesses respectives sont commandées électroniquement par une tête de sondage qui contrôle l'angle d'avance du fil de façon à assurer, autant que possible, que chaque tour de fil soit posé avec précision auprès du tour précédent.

Le coefficient d'encombrement considérablement amélioré des enroulements que garantit cette machine donne un rendement électromagnétique maximum. Ce coefficient est d'une importance particulière dans le cas de pièces comme les relais et les solénoïdes, du type miniature surtout. La conformation rectilinéaire beaucoup plus précise des enroulements bobinés par cette machine augmente les possibilités d'emploi de bobinages non-soutenus là où des formes de soutien étaient nécessaires auparavant.

EE 8751 *pour plus amples renseignements*

BRAYHEAD (ASCOT) LTD (167, 169)

Full View Works, Kennel Ride, Ascot, Berkshire

DISPOSITIFS D'ACCORD DE TÉLÉVISION

(Illustration à la page 236)

Le dispositif d'accord PB1 constitue un accessoire à 10 positions de grande efficacité et de construction compacte. Il permet la charge univoie ou multivoies, une gamme d'accords précis et le son à fréquence modulée sur une ou trois positions. Il offre un haut niveau de rendement en ce qui a trait à l'amplification, au glissement, au bruit, à la stabilité et à la forme de réponse.

Les dispositifs d'accord PB3 et PB4 sont le produit des efforts conjugués des sociétés Plessey et Brayhead. Ils ont le même encombrement réduit que le PB1, fournissant 14 voies, en ce qui concerne le PB3, et 18 voies pour le PB4, tous deux offrant une position supplémentaire permettant d'ajouter, s'il y a lieu, un raccord hyperfréquences.

Le branchement télévision-modulation de fréquence peut être effectué soit par un interrupteur plat fixé sur la rallonge de broche arrière soit par un interrupteur auto-nettoyeur à cames. Ce dernier est fourni en modèles pour accord modulation de fréquence univoie ou multivoies.

L'utilisateur a le choix de nombreux autres avantages comprenant:
filtre f.i. imprimé, accord de fréquence modulée univoie ou multivoies, composantes d'isolement incorporées, circuit d'accord précis pour bande large ou étroite, bobines d'accord d'antenne réglables, circuit d'entrée à couplage inductif, circuit de sortie f.i. à double accord, charge univoie ou multivoies, circuit de sortie f.i. à double fréquence, commutation de sensibilité incorporée, commutation d'antenne incorporée.

On peut, par ailleurs, employer le tube PCC89 ou la combinaison 30L15/30C15.

En plus des voies normales anglaises, ces dispositifs peuvent être pourvus de voies pour les systèmes de la Commission fédérale des communications (E.U.) ou du Comité consultatif international des radio-communications, ainsi que pour toutes les autres voies spéciales utilisées à l'étranger.

EE 8752 *pour plus amples renseignements*

FOIRE INTERNATIONALE DE HANOVRE

"Electronic Engineering" occupera le Stand 2, dans la Salle 13, à la Foire Internationale de Hanovre, du 26 avril au 13 mai 1959, et les visiteurs y seront les bienvenus.

BRITISH PHYSICAL LABORATORIES (88)

Houseboat Works, Radlett, Hertfordshire

PONT D'IMPÉDANCES UNIVERSEL

(Illustration à la page 236)

Ce pont d'impédances universel est un instrument pouvant effectuer une série étendue de mesures de valeurs de résistance d'inductance et de capacité.

Lorsqu'il est employé pour les mesures de résistance, il est relié, sous forme de pont Wheatstone, à une tension continue de pont intérieure. Afin d'empêcher l'auto-échauffement, en particulier de résistances au charbon de faible puissance, la tension de pont a été réduite à une valeur qui empêche—quelle que soit la mesure effectuée—la charge de la résistance sous examen de dépasser 16 mW. La tension déséquilibrée est amplifiée par un amplificateur écrêteur, l'indication s'effectuant sur un enregistreur de 10 cm muni d'une protection automatique de surcharge.

En alliant une résistance à variation continue à une résistance de décades, une longueur d'échelle exceptionnellement longue (18,3 mm) peut être obtenue. Ce procédé est employé pour les mesures de résistance de capacité et d'inductance et il donne une grande précision de lecture.

Lorsque l'instrument est employé comme pont de capacité, il est disposé sous forme de pont de Saute modifié, alimenté par un oscillateur intérieur en tension de pont de 1 kHz. La perte diélectrique est mesurée par une seconde commande variable combinée avec un interrupteur à cinq directions. Il est également possible de brancher une source de courant continu externe afin de pouvoir appliquer une tension continue superposée à la capacité sous essai. La détection de zéro s'effectue au moyen d'un amplificateur-détecteur accordé suivi par un indicateur de 10 cm, muni d'une protection de surcharge automatique.

Pour mesurer l'inductance, l'appareil est changé en pont de Hay, l'étalon réactif consistant en un condensateur au mica argenté. Le contrôle est le même que pour l'équilibre de capacité et de facteur de puissance. En branchant une source extérieure d'alimentation en continu, on peut faire passer une tension par le bobinage sous examen.

EE 8753 *pour plus amples renseignements*

BURNDEPT LTD (121)

Erith, Kent

DÉTECTEUR DE CONTENU POUR BOITES EN CARTON

(Illustration à la page 237)

Le Détecteur de contenu BURNDEPT Type BE.251 utilise un interrupteur de proximité qui détecte avec grande rapidité le contenu métallique ou non-métallique de boîtes en carton, à condition que ce contenu présente à la sonde de détection un changement de capacité électrique suffisamment important. L'appareillage de détection du contenu de boîtes en carton comprend:

(a) Un détecteur de contenu BE 251
(b) Une tête d'éjection de détecteur de contenu BE 257
(c) Deux connecteurs de détecteur de contenu BE 247
(d) Une tête de sondage de carton BE 258
(e) Une tête de sondage de contenu BE 243

Les accessoires (d), (e) et, parfois, (b) sont construits sur commande pour répondre à chaque application particulière, c'est à dire en fonction des dimensions des cartons employés.

La tête d'éjection et les sondes de détection sont généralement montées près de la courroie de transmission, tandis que les autres éléments sont montés à l'intérieur d'une boîte à instruments pouvant être fixée sur un mur voisin.

Fondamentalement, les deux interrupteurs de proximité servent de détecteurs tandis que l'élément de contrôle coordonne leurs fonctions avec l'action de la tête d'éjection. L''un des interrupteurs de proximité vérifie (au moyen de sa sonde de détection) si une boîte est vide ou pleine, cependant que l'autre détermine s'il y a des boîtes sur la chaîne de transmission. Ce dernier perfectionnement a été introduit afin d'empêcher que la tête d'éjection ne fonctionne inutilement lorsque la chaîne est complètement dégarnie de boîtes en raison, par exemple, de l'espacement ou d'une panne de production.

Si l'on veut répondre à tous les systèmes de production, il faut placer les deux sondes de détection l'une au dessus de l'autre. En cas d'emploi de systèmes à courroies de transmission, la sonde inférieure (tête de sondage de carton) doit être insérée à l'intérieur de la ligne de débit de façon à ce que les articles soient pressés contre la surface.

Le détecteur BE 251 peut détecter une gamme variée de produits liquides ou solides, en poudre ou en pâte. Aliments, médicaments, poudres détergentes, boulettes de métal, etc. peuvent tous être vérifiés avant ou après la fermeture du carton.

Il arrive parfois que certains articles ne peuvent être détectés par aucune autre méthode. Celà est vrai, par exemple, des cartons contenant des films, qui ne pourraient être examinés par une méthode nucléaire car celà causerait le voilement du film. Seule la méthode capacitive peut être employée en ce cas, pour s'assurer si un carton est plein ou vide. Enfin, le BE251 peut, en certains cas, être employé pour déterminer si des cartons sont correctement remplis à un niveau fixé d'avance.

EE 8754 *pour plus amples renseignements*

CATHODEON CRYSTALS LTD (7)

Linton, Cambridgeshire

FOUR À CRISTAL MINIATURE

(Illustration à la page 237)

Spécialement étudié pour le contrôle précis de température des cristaux du type 2M, le four à cristal miniature CATHODEON convient particulièrement aux systèmes radiotéléphoniques mobiles à voies serrées où l'espace est fort restreint mais dont le contrôle de fréquence précis est, néanmoins, essentiel. Une stabilité de fréquence supérieure à ±0005%, dans une gamme étendue de températures, peut être obtenue aisément. Le four se distingue, en outre, par son encombrement réduit, sa consommation électrique minime et son chauffage rapide.

Prévu pour une alimentation de 6 ou 12 volts, l'appareil a une consommation électrique de 4,6 watts et il peut être réglé pour des températures de 75°C, 80°C et 85°C. La durée de chauffage est de moins de 5 minutes entre +20°C et +85°C. Enfin, il ne pèse que 43kg.

EE 8755 *pour plus amples renseignements*

CRISTAUX SUBMINIATURE

Un nouveau type de cristaux subminiature à prises vient d'être lancé sous la référence 2MM. Ces éléments sont particulièrement indiqués pour l'emploi avec les transistors et autres composantes miniature. Ils sont prévus pour une gamme de fréquences entre 5 et 75 MHz (de 5 MHz fondamentaux à 20 MHz et de 15 MHz résonance série à 75 MHz), avec un rendement de ±0,005%, de −55°C à 105°C.

EE 8756 *pour plus amples renseignements*

DAWE INSTRUMENTS LTD (66)

99 Uxbridge Road, Ealing, London W.5

VOLTMÈTRE ÉLECTRONIQUE INDIQUANT LA VALEUR EFFICACE RÉELLE

(Illustration à la page 237)

Au contraire des voltmètres électroniques courants, qui sont des instruments de lecture des valeurs moyennes et de crête, étalonnés pour la lecture des valeurs efficaces, le Voltmètre Type 612 mesure avec précision la valeur efficace réelle des formes d'ondes non sinusoïdales. Il permet donc d'obtenir des indications sûres de la valeur efficace de tensions ou de courants se rapportant à des dispositifs ou à des circuits non-linéaires. Treize gammes de tensions sont prévues avec des déviations totales de 300 μV, 1 mV, 3 mV, 100 mV, 300 mV, 1 V, 3 V, 10 V, 30 V, 100 V et 300 V. La réponse en fréquence est de ±3% entre 15 Hz et 150 kHz, et de ±5% entre 5 Hz et 500 kHz.

EE 8757 *pour plus amples renseignements*

OSCILLATEUR B.F. DE GRANDE PUISSANCE

(Illustration à la page 237)

Cet oscillateur est du type résistance-capacité accordée et sa gamme de fréquences s'étend de 20 Hz à 20 kHz. Il se caractérise par sa grande puissance de sortie (6 watts), s'accordant avec une impédance de charge de 15 ohms ou de 60 ohms, et lui permettant d'assurer directement le fonctionnement de haut-parleurs ou de petits générateurs de vibrations. Sa précision est de ±1% dans la gamme de 200 Hz à 20 kHz. Dans les autres gammes, elle est de ±2%.

EE 8758 *pour plus amples renseignements*

ELECTROTHERMAL ENGINEERING LTD (135)

270 Neville Road, London E.7

EQUIPEMENT POUR BRASAGE DE PRÉCISION

(Illustration à la page 237)

Cet équipement constitue un ensemble portatif compact et d'un fonctionnement aisé, destiné à l'emploi avec des alliages de brasage à base d'argent. Il est particulièrement utile pour le brasage des fils de résistances, en nickel-chrome ou en alliages analogues, c'est à dire lorsque la soudure tendre est difficile à utiliser ou lorsqu'elle a un point de fusion trop bas pour la tâche envisagée ou, encore, lorsque la soudure par points donne une couture trop faible.

Quoique prévu principalement pour le travail de précision (diamètre à partir de 0.025 mm), cet équipement peut également souder des fils atteignant jusqu'à 3,25 mm d'épaisseur.

La consommation de courant de l'équipement dépend du réglage de contrôle. Le maximum est d'environ 500 watts. La soudure est chauffée au moyen d'un élément usable mais pouvant, cependant, être facilement remplacé. Des récipients à flux décapants et à solutions pour enlever ces derniers sont fixés de façon commode pour l'opérateur, les solutions étant, bien entendu, chauffées. L'alimentation de l'équipment est prévue pour secteur alternatif, de 200 à 250 volts, sans réglage, ainsi que pour d'autres tensions alternatives sur commande spéciale.

EE 8759 *pour plus amples renseignements*

FORTIPHONE LTD (152)

92 Middlesex Street, London E.1

POTENTIOMÈTRES SUBMINIATURE

Le Potentiomètre subminiature type VS32 à interrupteur unipolaire est de dimensions extrêmement réduites et il a été réalisé à l'usage des amplificateurs pour sourds (appareils auditifs). Conçu pour être fixe frontalement ou de côté, il est tenu par une vis et un écrou. Son diamètre total est de 8 mm et sa largeur est de 3,8 mm. Ses valeurs ohmiques vont de 1 kilohm à 2 mégohms et sa puissance nominale est de 50 mW.

Le Potentiomètre préréglé type PSR est destiné à l'emploi dans les circuits bobinés conventionnels ou dans les circuits imprimés. Son diamètre total est de 11,4 mm et sa largeur est de 2 mm. Les valeurs ohmiques et la puissance nominale sont identiques à celles du type VS32.

Le type PSS est un potentiomètre glissant préréglé de lignes et de construction simples dont la longueur totale est de 28 mm et la largeur totale de 7,2 mm. Sa gamme ohmique va de 1 kilohm à 3 mégohms et sa puissance nominale est de 50 mW.

EE 8760 *pour plus amples renseignements*

GEO. L. SCOTT & CO. LTD (133)

Cromwell Road, Ellesmere Port, Cheshire

NOYAUX ET TÔLES DÉCOUPÉES POUR TRANSFORMATEURS

Cette société vient d'ajouter à sa production, une nouvelle gamme de tôles subminiature pour transformateurs dont les dimensions hors-tout vont de 23 mm × 19 mm à 9,5 mm × 9,5 mm. Ces tôles sont destinées à l'usage des circuits de transistors, où l'espace est fort restreint.

Cette production comprend également des noyaux toroïdaux à base de fer-nickel, réalisés par traitement du fer en poudre. Ces noyaux sont fournis dans les normes RCL 193, sur bobines céramique pour hautes températures, pouvant recevoir directement les enroulements. Les caractéristiques constantes et le grand rendement qu'assurent ces alliages se vérifient lorsqu'on a besoin, par exemple, de paires accordées pour des amplificateurs magnétiques. Des noyaux toroïdaux à base de matériaux de fer au silicium, de 3,33 mm, 0,1 mm ou 0,05 mm d'épaisseur, peuvent également être fournis dans les normes RCL 193 ou dans celles des clients, pour des sources de puissance de basse et haute fréquence.

Enfin, des bandes à commutation rapide et grande perméabilité, dont l'épaisseur peut se réduire à 0,008 mm, sont produites sur bobines céramique pour matrices de rupture et autres dispositifs magnétiques de grand rendement.

EE 8761 *pour plus amples renseignements*

HINCHLEY ENGINEERING CO. LTD (91)

Pans Lane, Devizes, Wiltshire

TRANSFORMATEURS D'ISOLEMENT

(Illustration à la page 238)

En plus d'une gamme variée de transformateurs pour appareils de radio et de télévision, cette société exposera également un choix classique de transformateurs d'isolement et d'auto-transformateurs pour laboratoires, ateliers et services de dépannage. Ces transformateurs sont tous munis de sélecteurs de tension de sortie à fusibles rapidement remplaçables, et peuvent être fournis sur demande avec des bornes Mycalex à action instantanée.

Deux versions de la "boîte de sûreté" Hinchley constituent des éléments d'un intérêt spécial en raison du nombre croissant d'appareils électroménagers employés dans les ateliers et les jardins. Ce sont des transformateurs d'isolement portatifs, renfermés dans de robustes coffrets d'acier (le rapport d'isolement est de 1:1 ou, de préférence, une sortie de 110 V ou 55 V avec point central à la terre).

Ces transformateurs ont été spécialement étudiés pour la production en grosses quantités, de sorte que le coût final au consommateur fera de la sécurité une nécessité plus économique.

EE 8762 *pour plus amples renseignements*

JAMES A. JOBLING & CO. LTD (158)

Wear Glass Works, Sunderland

RÉSISTANCES À L'OXYDE MÉTALLIQUE

Ces résistances de haute précision et de grande stabilité sont faites de barres "Pyrex", revêtues d'une couche mince d'oxyde métallique. Le noyau et la couche d'oxyde métallique gardent toute leur imperméabilité et leur stabilité dans les conditions les plus défavorables et leur niveau de bruit est extrêmement bas.

La tolérance normale de ces résistances est de 1% et elles sont fabriquées sur commande, de façon à répondre aux besoins précis du client plutôt qu'à certaines préférences générales, dans les gammes de 100 ohms à 250 kilohms et pour une puissance de 0,5 watts à 2 watts.

EE 8763 *pour plus amples renseignements*

LABGEAR LTD (110)

Willow Place, Cambridge

CONTRÔLEUR D'ISOLEMENT NON-DESTRUCTIF

(Illustration à la page 238)

Cet appareil a été conçu pour les essais d'isolement de circuits et de composantes tels que les condensateurs faibles.

Le modèle standard assure des tensions de contrôle en alternatif de 250 V, 500 V, 1 kV, 1,5 kV et 2 kV. Une des caractéristiques de l'appareil est qu'il permet de contrôler une composante fonctionnant à une tension supérieure à sa tension de service normale. En cas de panne, la composante n'est nullement endommagée et peut être utilisée à nouveau à un potentiel plus faible. Le contrôle de composantes et d'accessoires peut donc être effectué sans risque d'endommagement.

EE 8764 *pour plus amples renseignements*

LUSTRAPHONE LTD (111)

St. George's Works, Regents Park Road, N.W.1

MICROPHONE STÉRÉOPHONIQUE

(Illustration à la page 238)

Cet appareil microphonique se compose de deux microphones à rubans identiques, montés en ligne verticalement. La partie supérieure de l'élément, qui contient l'un des microphones, est construite de manière à pouvoir être tournée continûment sur un angle de 100° environ, à partir de la position "en ligne," permettant ainsi d'obtenir le meilleur réglage dans n'importe quelle éventualité (de transmission stéréophonique, enregistrement, radiodiffusion, renforcement du son, etc.). La partie inférieure est montée de manière fixe.

L'appareil est relié à un élément de commutation au moyen d'un connecteur à vis à quatre directions. L'avantage de cette méthode est qu'elle permet de détacher instantanément la tête sensible du microphone et de la ranger en lieu sûr, sans, toutefois, déranger le montage et les connexions vers l'installation principale.

Un interrupteur de conception spéciale est prévu pour trois positions, à savoir:

1. Position "arrêt" au centre.
2. Branchement stéréophonique à l'extrémité d'un des côtés.
3. Position à l'autre extrémité de l'interrupteur, reliant en série les deux éléments à rubans et changeant ainsi l'appareil en un microphone à ruban de vitesse de haute sensibilité et à caractéristique polaire "8".

Un interrupteur déphaseur, fixé à l'arrière de l'élément de commutation, permet l'inversion de phase instantanée, de manière à assurer la phase voulue sur n'importe quel équipement.

Lorsque le Microphone Stéréophonique est à la susdite position 3, et que les deux éléments sont "en ligne," on peut le changer, au moyen de l'interrupteur déphaseur, en un microphone éliminateur de bruit fort efficace, ne répondant à une source de bruit qu'à une distance rapprochée.

Conjointement avec les positions alternatives de l'interrupteur déphaseur, on obtient, par le positionnement intermédiaire correspondant des éléments microphoniques, divers effets directionnels fort utiles dans des conditions acoustiques défavorables.

L'impédance de chacun des éléments microphoniques dans le modèle standard est de 20 ohms et l'impédance totale de l'appareil composé, branché sur la position de série, est de 40 ohms.

EE 8765 *pour plus amples renseignements*

MC MURDO INSTRUMENT CO. LTD (32)

Victoria Works, Ashtead, Surrey

SUPPORTS DE LAMPE ET CONNECTEURS

Une nouvelle gamme de supports de lampes "petite plume," à moulage vierge simple du type B9A ou B7G, pour circuits imprimés et câblages classiques, est présenté par cette société.

Le moulage est conçu pour recevoir une griffe à ressort ordinaire, faisant partie intégrante de la monture métallique du support de lampe. En outre, les lampes peuvent être blindées au moyen d'écrans métalliques peu coûteux, épousant étroitement l'enveloppe de la lampe. Ces écrans sont tenus et mis à la terre électriquement par la griffe à ressort.

La gamme des fiches et douilles PAKONECTOR a été augmentée par la réalisation d'un modèle à 8 directions, de la même grandeur à peu près que le support de lampe B7G. On trouve les mêmes avantages de raccordement de câbles dans le modèle à 8 directions que dans les versions originales à 18 directions.

EE 8766 *pour plus amples renseignements*

MULLARD LTD (77/67)

Mullard House, Torrington Place, London W.C.1

TUBES À RAYONS CATHODIQUES DE 110°

Des tubes cathodiques de 110° à écrans de 534 mm et 432 mm sont produits par cette société sous les références de modèles AW53-88 et AW43-88 respectivement.

Ces nouveaux tubes sont sensiblement plus courts que les tubes de diamètre égal d'usage courant et ils permettent donc de réduire la profondeur des récepteurs de télévision.

Le tube de 110° à écran de 534 mm est plus court de 216 mm que le tube de même écran de 70°, et plus court de 127 mm que celui de 90°. Quant au tube de 110° à écran de 432 mm, il est plus court de 165 mm et de 76 mm que les tubes à écrans équivalents de 70° et 90° respectivement.

La substance luminescente de grande efficacité des nouveaux tubes est à fond métallique, assurant ainsi une brillance et un contraste parfaits dans les conditions de forte lumière ambiante. Leur focalisation est électrostatique et ils comportent un canon à électrons du type "droit", ce qui rend superflu l'aimant piège à ions.

EE 8767 *pour plus amples renseignements*

PAINTON & CO. LTD (29)

Kingsthorpe, Northampton

RÉSISTANCES MINIATURE À HAUTE STABILITÉ

Ces résistances au charbon à haute stabilité, du type 70, sont de dimensions extrêmement réduites, ne mesurant, à peu près, que 9,5 mm × 2,5 mm. Elles sont donc indiquées pour l'emploi dans des équipements transistorisés miniature et, en particulier, pour les plaques a circuits imprimés. Les conducteurs sont préétamés afin de faciliter leur brasage par immersion. La gamme de résistances s'étend à toutes les valeurs standard, de 5 à 250 kΩ, avec une tolérance de 5%, mais des tolérances de 1 et de 2% peuvent, si nécessaire, être prévues sur presque toute la gamme.

EE 8768 *pour plus amples renseignements*

CONNECTEURS DE CIRCUITS IMPRIMÉS

(Illustration à la page 239)

Un nombre de connecteurs de circuits imprimés vient d'être mis sur le marché. Il comprend une fiche à 15 directions et une prise à goujons d'acier. De même que tous les connecteurs PAINTON pour circuits imprimés, les contacts sont sur module de 2.54 mm.

La production Painton comporte également un connecteur coudé comprenant jusqu'à 60 contacts, en trois groupes de 20 avec centres de 2.54 mm. Ces contacts sont d'un modèle spécial breveté et les essais ont démontré qu'ils permettent une insertion sur plaque de très longue durée.

EE 8769 *pour plus amples renseignements*

SUPPORTS DE LAMPES À ANGLE DROIT

(Illustration à la page 239)

Ces composantes ont été développées afin de permettre le montage horizontal des lampes, facilitant ainsi la production d'assemblages de circuits imprimés plus compacts et simplifiant le dispositif de source froide.

Les contacts sont en cuivre au glucinium argenté et leur longueur est telle qu'elle s'adapte aux épaisseurs de circuits imprimés les plus communément employées. Les pressions correctes d'insertion et de retrait sont maintenues avec le minimum de résistance de contact.

EE 8770 *pour plus amples renseignements*

STEATITE AND PORCELAIN PRODUCTS LTD (76)

Stourport-on-Severn, Worcestershire

JOINTS À BORNES

(Illustration à la page 239)

Selon l'annonce préliminaire de l'année dernière, la nouvelle gamme de joints à bornes se conforme à la Spécification militaire RCL 331. Ces joints sont en céramique à base étendue d'alumine et les tiges sont fermement fixées en place par fusion. Des bandes métallisées extérieures, réalisées par le procédé de métallisation de nickel de qualité agréée, sont fournies conjointement. Elles permettent le brasage tendre direct dans les enveloppes métalliques pour toutes les applications de fermeture hermétique.

Les gammes standard de joints métallisés au nickel de la qualité agréée dans la catégorie H1, sont toujours livrables du stock et un service spécial chargé de la fabrication d'échantillons fournit, sur court préavis, des prototypes de nouveaux modèles. D'autre part, on a considérablement développé les services d'assemblage des produits en céramique métallisée, avec des ferrures appropriées, afin de répondre à une variété étendue d'applications allant des cosses miniature aux grandes bornes haute tension des condensateurs.

EE 8771 *pour plus amples renseignements*

SWIFT LEVICK & SONS LTD (156)

Clarence Steel Works, Sheffield 4

GRANDE AIMANTS PERMANENTS

(Illustration à la page 240)

Cette société vient d'ajouter à sa production les grands aimants stables à haute homogénéité pour les applications de résonance magnétique nucléaire. Notre gravure montre un de ces aimants, qui a été étudié et produit pour les travaux de recherches de l'Université de Leeds. Cet aimant pèse 2350 kg. et son encombrement est de 1,21 m × 1,07 m × 0,61 m. Il est excité de façon permanente par l'alliage Columax, qui constitue la matière pour aimants la plus stable et la plus puissante qui puisse être produite commercialement en n'importe quelle partie du monde. L'aimant a une puissance du champ d'entrefer de 9700 Oersteds sur un volume d'entrefer de 36 mm de long × 254 mm de diamètre. L'uniformité du champ est supérieure à 1 partie dans 10^6 sur un volume central d'environ 1 cm³. L'une des pièces polaires en fer est pourvue d'un dispositif d'ajustement mécanique spécial, permettant un mouvement réduit de la face polaire sur n'importe quel plan afin de pouvoir effectuer un réglage final uniforme de la puissance du champ d'entrefer.

Un modèle d'un sixième de la grandeur de cet électroaimant sera exposé au Salon.

EE 8772 *pour plus amples renseignements*

TELEPHONE MANUFACTURING CO. LTD (21)

Hollingsworth Works, London S.E.21

RELAIS MULTIPLE

(Illustration à la page 240)

Le nouveau relais multiple à 10 directions TMC sera exposé au Salon. De nombreux procédés entièrement nouveaux ont été utilisés dans sa fabrication, dont le résultat a été d'en réduire le prix en augmentant son efficacité. Le relais multiple TMC emploie une culasse commune faisant partie intégrante des supports de fixation et il est destiné à remplacer les relais de type téléphonique dans les circuits de contrôle où le coût des éléments joue un rôle primordial.

Les ressorts de contact en bronze phosphoreux, avec contacts bimétalliques d'argent, selon la technique dite des rayons croisés, ainsi que la réglage mécanique automatique et le fonctionnement, assurent une pression de contact constante qu'on ne trouvait jusqu'ici que sur les relais coûteux de précision. Un circuit magnétique efficace empêche virtuellement toute interaction magnétique entre les relais, tandis que les supports de fixation à culasse commune réduisent grandement le nombre des pièces et rendent inutiles les supports séparés.

Le prix de ce relais multiple est tel qu'on constatera souvent qu'il est plus économique d'installer un élément à 10 directions qu'un moindre nombre de relais conventionnels.

EE 8773 *pour plus amples renseignements*

WHITELEY ELECTRICAL RADIO CO. LTD (81)

Radio Works, Victoria Street, Mansfield, Nottinghamshire

CONTRÔLEUR DE CABLES

(Illustration à la page 240)

Cet appareil de contrôle localise, à quelques millimètres près, l'endroit d'une coupure ou d'un court-circuit dans les câbles armés à plusieurs conducteurs. Ces derniers peuvent être, ensuite, réparés avec un minimum de dérangement de l'armure.

Le principe de fonctionnement est comme suit:

En cas de conducteur brisé, une tension de fréquence acoustique est appliquée entre l'une des extrémités du conducteur et la terre. Le champ électrostatique entre la partie sous tension du conducteur et la terre peut être, ensuite, détecté en faisant passer une sonde capacitive le long du câble. La sonde est reliée à l'entrée d'un amplificateur à transistors portatif qui alimente un casque d'écoute, donnant ainsi une indication audible de l'endroit de la rupture. Tous les autres conducteurs, à l'exception de celui que l'on contrôle, doivent être mis à la terre.

Lorsqu'on veut repérer un court-circuit entre des conducteurs, on emploie une sonde inductive au lieu d'une sonde capacitive. L'oscillateur ne transmettra un signal au conducteur que jusqu'au point précis où le court-circuit se produit et la sonde détectera le champ magnétique jusqu'à ce point.

Le tableau de contrôle est muni de commutateurs rotatifs pouvant choisir des conducteurs particuliers à l'intérieur du câble. Les conducteurs autres que celui qui a été choisi sont mis à la terre. Des prises sont pourvues pour recevoir les fiches des câbles. Un ohmmètre du type à lampe est fourni pour pouvoir s'assurer au préalable de l'état d'un conducteur.

L'oscillateur à lampe fonctionne sur pile; il a une puissance de sortie de 10 mW et une impédance de sortie de 300 ohms. Sa fréquence est de 1 kHz à ±2,5%. Le type d'oscillation peut être soit continu soit interrompu, selon le réglage du potentiomètre.

L'amplificateur à transistors fonctionne sur pile de 4,5 volts renfermée dans le boîtier et il a un gain total de 70 décibels lorsqu'il est relié à une charge non inductive de 300 ohms. La sonde inductive est accordée à kHz ±2,5%.

EE 8774 *pour plus amples renseignements*

THE ZENITH ELECTRIC CO. LTD (8)

Zenith Works, Villiers Road, Willesden Green, N.W.2

TRANSFORMATEUR RÉGULATEUR

(Illustration à la page 240)

Un nouveau modèle de transformateur, le "VARIAC 50-B/2B", à double sortie, sera exposé au Salon cette année. Il a été étudié en vue d'une alimentation de 230 volts, 50 Hz, assurant une puissance de sortie variant entre 0 et 270 volts et une intensité nominale maximum de 20 A, ce qui constitue le courant total pouvant être tiré des deux balais, mais la charge peut être, évidemment, répartie de manière à pouvoir répondre, par exemple, à une application exigeant 12 A —Sortie Uo. 1 et 8 A—Sortie No. 2.

Les transformateurs VARIAC à deux balais et à double sortie existent depuis longtemps déjà dans la série "100" et "V30H", ainsi qu'en format réduit dans la gamme toroïdale de la série "200".

Le modèle réduit a maintenant été modifié en employant certaines pièces de la série "V5", ce qui a donné un élément bien plus compact.

EE 8775 *pour plus amples renseignements*

CONTRÔLEUR DE CIRCUITS DE RETOUR PAR LA TERRE

Cet appareil est logé dans un meuble métallique muni d'une manette de support et d'une pochette pour câble d'entrée et fiche. Il a été étudié en vue de pouvoir exécuter intégralement tous les essais exigés par le Règlement No. 507 de l'INSTITUTE OF ELECTRICAL ENGINEERS, ainsi que pour fonctionner sur courant monophasé ordinaire d'une puissance de sortie de 25 A, à n'importe quelle tension entre 0 et 40 volts.

Un transformateur abaisseur bifilaire ZENITH est employé avec un VARIAC à travers le primaire et un voltmètre de redresseur est fourni, de même qu'un ampèremètre. En outre, une prise de courant à trois broches de 13 A, ainsi qu'une prise à trois broches de 15 A, se trouvent à l'avant du coffret pour le contrôle du noyau de mise à la terre des appareils portatifs. La tension de 40 volts est prévue pour la plupart des essais mais celle de 6 volts est également comprise car elle convient mieux éventuellement aux petits circuits bouclés comme ceux des outils portatifs ou autre matériels d'outillage analogues.

Enfin, un coupleur est prévu pour le contrôle d'appareils fixes, pour lequel la tension de 6 volts est employée, car le fil de masse est relativement court et ne dépasse pas 0,1 ohm, ce qui représente environ 3 volts.

EE 8776 *pour plus amples renseignements*

ERRATA :

Nous nous excusons auprès de nos lecteurs des erreurs qui se sont glissées dans les indications des vitesses d'avance de la bande de l'Enregistreur à bande magnétique pour calculateurs électroniques, construit par la Société Epsylon Industries Limited, dans la description que nous en avons donné à la page 193 de notre numéro de mars 1959. Ces vitesses sont en réalité comme suit:

Vitesse actuelle de la bande: 254 cm/sec.

Vitesse future de la bande: 508 cm/sec.

Vitesse de bobinage accéléré: **1016 cm/sec.**

Résumés des Principaux Articles

Potentiomètres linéaires à plusieurs prises et à sorties chargées par K. C. Garner

Résumé de l'article aux pages 192 à 199

L'auteur analyse les potentiomètres linéaires à plusieurs prises, du type à résistances de shunt reliées entre prises intermédiaires adjacentes faisant face à une résistance de charge de sortie.

En utilisant le rapport d'un réseau généralisé avec des substitutions numériques appropriées, obtenues d'un dispositif à potentiomètres dont l'auteur traite en particulier, l'évaluation devient un procédé simple et systématique.

L'auteur propose en conclusion une méthode pour la fabrication de potentiomètres produisant des fonctions non-linéaires.

L'analyse des bruits de chocs par F. M. Savage

Résumé de l'article aux pages 200 à 203

Les instruments pour la mesure et l'analyse de bruits persistants ont permis d'établir des critères de recherches en vue de prévenir la surdité chez les personnes exposées à des bruits continus. Le perfectionnement des analyseurs de bruits de chocs devraient fournir des données pour l'étude de la

réduction des bruits de chocs les plus dangereux pour l'ouïe. Cet article décrit un petit appareil portatif pour la mesure de bruits d'impacts. Cet appareil peut également servir à d'autres usages, tels que la mesure d'accélérations de pointe, lorsqu'on l'utilise avec un accéléromètre adéquat.

Amplificateur à transistors pour le réenregistrement de bandes magnétiques et rouleaux. par A. E. Bachmann.

Résumé de l'article aux pages 213 à 217

On emploie fréquemment les bandes magnétiques et les rouleaux comme éléments enregistreurs ou émetteurs contenant les données sous forme numérale. Ces renseignements peuvent être obtenus à nouveau en réenregistrant la bande ou le rouleau voulu au moyen d'une tête magnétique et d'un amplificateur. Cet article examine les problèmes que pose la réalisation d'un pareil amplificateur en n'employant que des transistors. Il décrit, en particulier, un prototype d'amplificateur dont on vient d'effectuer les essais et qui fonctionne à partir d'une tête de réenregistreemnt dont la tension de sortie à circuit ouvert dépasse 2 mV. Cet amplificateur est capable d'actionner en parallèle trois multivibrateurs bistables dans une gamme de fréquences de 2 à 400 kHz. et une gamme de températures de 0°C à + 70°C. La durée de montée de l'impulsion de sortie est de moins d'une microseconde.

Un oscillateur simple de très basse fréquence par J. F. Young

Résumé de l'article aux pages 218 à 220

Les méthodes usuelles de stabilisation d'amplitude d'oscillateurs résistance-capacité sont des méthodes sensibles aux fréquences. Dans l'oscillateur simple décrit dans cet article, une diode Zéner limite l'amplitude, de sorte que les harmoniques qui en résultent sont filtrés par un circuit sélectif, qui contrôle également la fréquence des oscillations. On peut donc obtenir une forme d'onde tolérable, même à de très basses fréquences.

Une méthode stroboscopique pour effectuer des mesures de réponse de fréquence sur des petits dispositifs électromécaniques par M. Shepherdson et R. Walters

Résumé de l'article aux pages 220 à 221

Cet article décrit une méthode permettant d'effectuer des mesures de réponse de fréquence sur des petits dispositifs électromécaniques utilisant un microscope mobile et une lampe stroboscopique. En dehors d'un simple amplificateur d'impulsions et d'un réseau de déphasage que les auteurs décrivent, l'attirail nécessaire à cette méthode se trouve généralement dans la plupart des laboratoires électriques bien outillés. La méthode n'implique pas le chargement de l'accessoire d'essai et permet d'effectuer des mesures d'amplitude bien au dessous de 0,025 mm.

Alimentation stabilisée par D. J. Collins et J. E. Smith

Résumé de l'article aux pages 222 à 226

L'alimentation accordée ou stabilisée en courant continu est nécessaire à divers usages. Le sauteurs examinent succinctement les différents types de régulateurs, décrivant, ensuite, plus en détail, le stabilisateur de circuit bouclé. Les problèmes de fabrication que pose ce genre de dispositif, sont également étudiés, tout particulièrement en ce qui concerne la stabilité du circuit fermé. Un dessin détaillé d'un stabilisateur-type figure à l'appendice.

La télémesure par l'application du principe du tube répartiteur à gaz par Yoshisuke Hatta

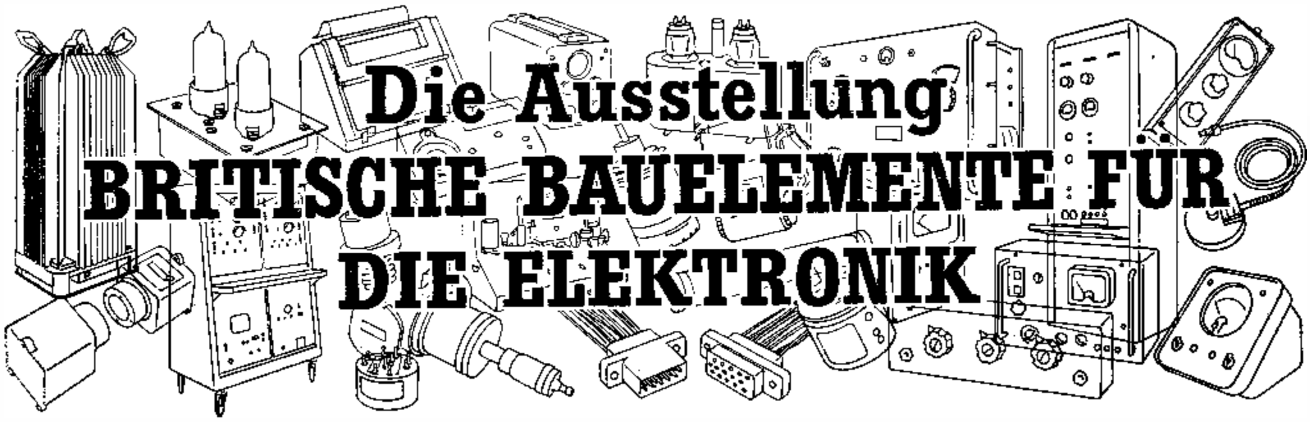
Résumé de l'article aux pages 227 à 229

L'auteur décrit une méthode de télémesure du déplacement mécanique par l'application du principe du tube répartiteur à gaz. Ce tube étant un élément électronique, il a un bon fonctionnement à action rapide et il est, donc, supérieur, à certains égards, aux systèmes électromagnétiques dont l'inertie est considérable.

Condensateurs au tantale à électrolyte solide. par R. Aries.

Résumé de l'article aux pages 230 à 231

Les condensateurs au tantale à électrolyte solide sont communément utilisés dans les appareils miniature. Cet article décrit brièvement leur procédé de fabrication ainsi que les résultats d'essais de durée et de propriétés thermiques.



Die Ausstellung BRITISCHE BAUELEMENTE FÜR DIE ELEKTRONIK

Beschreibungen, zusammengestellt auf Grund von Mitteilungen aus Herstellerkreisen, einer Auswahl von Geraten, die auf der Ausstellung des Verbands der Hersteller von Bauelementen für die Radio-und Elektronik-industrie im Grosvenor House und Park Lane House, London, vom 6.-9. April zur Schau gelangen werden. (*Die in Klammern angegebenen Zahlen und Buchstaben beziehen sich auf die Standnummern*)

(Ubersetzung der Seiten 236 bis 240)

AVO LTD (73)

Avocet House, 92-96 Vauxhall Bridge Road, London, S.W.1

SPULENWICKELMASCHINE

Eine neue Spulenwickelmaschine wird zur Ausstellung kommen, die in der Lage ist, beinahe vollkommen lagengewickelte Spulen ohne die Verwendung jeder Art von Verflechtungsvorrichtung zu wickeln. Die normale mechanische Kupplung zwischen Spindelstock und Traverse der Maschine wurde überflüssig gestaltet und beide werden durch unabhängige Motoren angetrieben, deren betreffende Drehzahlen mittels eines Sondenkopfs gesteuert werden, der den Drahtvorschubswinkel überwacht um zu gewährleisten, dass jede einzelne Windung sich genau, sofern praktisch möglich, an die vorhergehende anlegt. Der beträchtlich verbesserte Raumfaktor von mittels dieser Maschine hergestellten Spulen gewährleistet ein Höchstmass elektromagnetischer Leistung-ein Faktor der von besonderer Wichtigkeit für solche Bauelemente wie Relais und Solenoiden ist, insbesondere für Miniaturtypen, während die weitaus genauere rechteckige Profilierung von auf diese Weise gewickelten Spulen viele Möglichkeiten für die Verwendung von selbsttragenden Spulen eröffnet, wo früher Formvorrichtungen erforderlich waren.

EE 8751 *für weitere Einzelheiten*

BRAYHEAD (ASCOT) LTD (167, 169)

Full View Works, Kennel Ride, Ascot, Berks

FERNSEH-REVOLVERTUNER

(Abbildungen auf Seite 236)

Der Tuner PB. 1 ist ein Hochleistungs-10 Positionentuner von kompakter Konstruktion mit Vorrichtungen für Ein- und Mehrfachkanalbelastung, geschaltetem Feintuner-Bereich und Vorrichtung für Ein- oder Dreiposition—FM—Ton. Er bietet hohe Leistungsgüte in Bezug auf Verstärkung, Drift, Rauschen, Stabilität und Wiedergabeform.

Die Tuner PB. 3 und PB. 4 sind das Ergebnis der vereinten Bemühungen von Plessey und Brayhead. Sie besitzen dieselben kompakten Gesamtabmessungen wie der PB. 1, mit 14 Kanälen im PB. 3 und 18 Kanälen im PB. 4, wobei beide eine zusätzliche Position für den Zusatz einer UHF-Vorsatzvorrichtung freihalten.

Fernseh-FM-Schalten lässt sich entweder vermittels eines Flachschalters, der auf der rückwärtigen Spindelverlängerung angebracht ist, oder mittels eines nockengesteuerten Gleitschalters vornehmen, wobei der letztere in für Ein- oder Mehrfachkanal-FM-Abstimmen geeigneten Ausführungen zur Verfügung steht.

Der Benützer hat eine Auswahl von vielen Zusatzvorrichtungen, die die Anwendungsmöglichkeiten sehr vielseitig gestalten; unter diesen befinden sich:—

Gedrucktes ZF-Filter, Ein- oder Mehrfachkanal-FM-Abstimmen, eingebaute Isolierungsbaulemente, Schmal- oder Breitband-Feintuner, verstellbare Antennenspulen, induktiv-gekoppelte Eingangs-Schaltung, ZF-Ausgangsschaltung mit doppelter Abstimmung, Ein-oder Doppelkanalbelastung, ZF-Ausgangsschaltung mit Doppelfrequenz, eingebautes Empfindlichkeitsschalten, eingebautes Antennenschalten. Vorkehrung für die Verwendung der PCC. 89 Röhre, oder 30L15/30C15 Kombination ist gleichfalls getroffen.

Zusätzlich zu normalem Kanalisieren für die Britischen Inseln können diese Tuner mit Kanälen für F.C.C. oder C.C.I.R. Systeme, oder für weitere in andern Ländern erreichbare Sonderkanäle versehen werden.

EE 8752 *für weitere Einzelheiten*

BRITISH PHYSICAL LABORATORIES (88)

Houseboat Works, Radlett, Hertfordshire

UNIVERSALIMPEDANZBRÜCKE

(Abbildungen auf Seite 236)

Diese Universalimpedanzbrücke ist ein Allzweckgerät, mit einem weiten Messbereich von Widerstands-, Induktanz- und Kapazitanzwerten.

Bei Verwendung für Widerstandsmessung wird das Gerät als eine Wheatstone'sche Brücke mit einer internen Brückengleichspannungsspeisung angeschlossen. Um Eigenerwärmung zu verhindern, insbesondere bei Niederstrom-Kohlenstoffwiderständen wurde die Brückenspannung auf einen derartigen Wert reduziert, dass bei jeder beliebigen Messung die sich unter Prüfung befindliche Widerstandtslast einen Wert von 16 mW nicht übersteigt. Die unsymmetrische Spannung wird mittels eines Zerhackungsverstärkers verstärkt und die Anzeige gesschieht vermittels eines 102 mm Meters, der mit einer automatischen Überlastschutzvorrichtung ausgestattet ist.

Dank der Kombination eines kontinuierlich regelbaren Widerstands und eines Dekadenwiderstands wird eine ungewöhnlich grosse Skalenlänge, nämlich 18,3 m, erzielt. Diese Anornung wird zur Widerstands-, Kapazitanz- und Induktanzmessung verwendet, und ermöglicht eine erhöhte Genauigkeit.

Als Kapazitanzmessbrücke wird das Gerät in Form einer modifizierten de Saute-Brücke angeordnet und wird von einem internen Oszillator mit einer Brückenspannung von 1 kHz gespeist. Der dielektrische Verlust wird mit einer zweiten regelbaren Steuerung in Verbindung mit einem fünfgängigem Schalter gemessen. Vorkehrungen zum Anschluss einer externen Gleichstromquelle zum Anlegen einer überlagerten Gleichstromspannung an den sich unter Prüfung befindlichen Kondensator sind getroffen. Nullgleichrichtung wird mittels eines abgestimmten Verstärkerdetektors bewirkt, dem ein mit einer automatischen Überlastschutzvorrichtung ausgestatteter 102 mm Anzeiger nachgeschaltet ist.

Für die Messung von Induktanz wird das Gerät in eine Hay'sche Brücke umgewandelt, wobei der Reaktivstandard aus einem versilberten Mikakondensator besteht; Dieselben Steuerungen kommen zur Verwendung wie für Kapazitanz- und Kraftfaktorausgleich. Durch Anschliessen einer externen Gleichstromquelle kann ein Strom durch die siche unter Prüfung befindliche Spule geleitet werden.

EE 8753 *für weitere Einzelheiten*

BURNDEPT LTD (121)

Erith, Kent

KARTONINHALT-MONITOR

(Abbildungen auf Seite 237)

Der 'Burndept' Kartoninhalt-Monitor Typ B.E. 251 verwendet einen Annäherungsschalter, der bei hoher Geschwindigkeit den metallischen oder nicht-metallischen Inhalt Bestimmen kann, vorausgesetzt dass der kartonierte Gegenstand eine genügend grosse Veränderung des elektrischen kapazitiven Widerstands der Bestimmungssonde gegenüber aufweist.

Der Gerätesatz für den Kartoninhalt-Monitor umfasst:

(a) Einen Kartoninhalt-Monitor BE.251
(b) Einen Kartoninhalt-Monitor Auswerferkopf BE.257
(c) Zwei Kartoninhalt-Monitor Verbindungsstücke
(d) Einen Karton - Bestimmungskopf BE.258
(e) Einen Kartoninhalt-Bestimmungskopf BE.243

Die Posten (d), (e) und manchmal (b) müssen speziell konstruiert werden, um jeder betreffenden Anwendung gemäss der Grösse der zur Verwendung gelangenden Kartons zu entsprechen.

Der Auswerferkopf und die Bestimmungssonden werden in der Regel in der Nähe des Förderbands installiert, während die andern Einheiten im Innern eines gusseisernen Geräteschranks untergebracht sind, der an der nächststehenden Wand befestigt werden kann.

Grundsätzlich werden zwei Annäherungsschalter als Detektoren verwendet, während das Steuerungsaggregat ihre Leistung und das Betätigen des Auswerferkopfs koordiniert. Ein Annäherungsschalter bestimmt (über seine Bestimmungssonde) ob eine Schachtel voll oder leer ist, während der andere fesstellt, ob sich irgendwelche Schachteln auf dem Förderband befinden. Diese letztere Verfeinerung wurde einbegriffen, um etwaigen unnötigen Betrieb des Auswerferkopfs zu verhindern, im Falle sich überhaupt keine Schachteln auf dem Förderband befinden, wie z.B. durch Abstandseinteilung oder Fertigungsausfall.

Um vollständige Erfassung für alle Fertigungssysteme zu gewährleisten ist es erforderlich, die beiden Bestimmungssonden übereinander anzubringen. Wo Bandzubringsysteme zur Verwendung kommen, muss die untere Sonde (Kartonbestimmungssonde) innerhalb des Förderbands so eingebaut werden, dass die Gegenstände unter Druck über seine Oberfläche geführt werden.

Der Monitor BE.251 kann einen weiten Bereich von Erzeugnissen in fester, Pulver-, Pasten- oder flüssiger Form bestimmen. Nahrungsmittel, Drogen, Reinigungspulver, metallische Kügelchen usw. lassen sich alle sowohl vor als auch nach dem Verschliessen des Kartons prüfen. Manchmal lassen sich gewisse Gegenstände einfach nicht durch andere Methoden bestimmen, z.B. Kartons, in denen Filme verpackt sind, dürfen nicht vermittels einer Kernmethode untersucht werden, da sonst 'Bewölkung' des Films auftreten würde, und in diesem Fall darf lediglich die kapazitive Methode zur Anwendung kommen um zu bestimmen, ob ein Karton voll oder leur ist. In gewissen Fällen kann der BE.251 auch dazu verwendet werden zu bestimmen, ob die Kartons bis zu einer vorbestimmten Höhe richtig gefüllt.

EE 8754 *für weitere Einzelheiten*

CATHODEAN CRYSTALS LTD (7)

Linton, Cambridgeshire

MINIATUR-KRISTALLOFEN

(Abbildungen auf Seite 237)

Der speziell für genaue Temperaturregelung von Kristallen der Type 2M entworfene 'Cathodean' Miniatur-Kristallofen is besonders gut geeignet für fahrbare Nahkanal-Radiotelefonsysteme, wo Präzisions-Frequenzregelung unerlässlichist, die Platzfrage jedoch an erster Stelle steht. Frequenzstabilität, besser als $\pm 0{\cdot}0005\%$ innerhalb eines grossen Temperaturbereichs, lässt sich mit Leichtigkeit erzielen. Kompakte Konstruktion, niederer Kraftverbrauch, und schnelles Anheizen vom kalten Zustand sind eine der besonderen Eigenschaften.

Der Ofen betätigt sich bei einer Spannung von 6 V oder 12 V mit einem Kraftverbrauch von 4,6 W, und Temperatureinstellungen von 75°C, 80°C und 85°C stehen zur Verfügung. Die Aufheizzeit ist weniger als 5 Minuten von +20°C bis +85°C. Das Gewicht beträgt 43 g.

EE 8755 *für weitere Einzelheiten*

SUBMINIATURKRISTALLE

Eine neue Reihe von Stecker-Subminiaturkristall-Aggregaten wurde eingeführt. Diese Einheit, Typ 2MM, ist besonders zum Einsatz mit Transistoren und anderen Miniaturbauelementen geeignet. Sie stehen in einem Frequenzbereich von 5 bis 75 MHz (Grundfrequenz 5 bis 20 MHz und Reihenresonanzoberton 15 bis 75 MHz) mit einer Leistung von $\pm 0{\cdot}005\%$ von -55°C bis 105°C zur verfügung.

EE 8756 *für weitere Einzelheiten*

DAWE INSTRUMENTS LTD (66)

99 Uxbridge Road, Ealing, London, W.5

ABSOLUTEFFEKTIVWERT-RÖHRENVOLTMETER

(Abbildungen auf Seite 237)

Ungleich den gewöhnlichen Röhrenvoltmetern, die Geräte zum Ablesen von Durchnitts- oder Spitzenwerten sind, und deren Eichung für das Ablesen von Effektivwerten eingerichtet ist, misst das absoluteffektivwert Röhrenvoltmeter Typ 612 in genauer Weise den Effektivwert von Wellenformen, die anders als sinusförmig geartet sind. Genaue Effektivablesungen lassen sich für Spannungen und Ströme erzielen, die in Verbindung mit nicht-linearen Geräten oder Schaltungen zur Verwendung kommen. Dreizehn Spannungsbereiche stehen zur Verfügung mit Vollausschlägen von 300 μV, 1 mV, 3 mV, 100 mV, 1 V, 3 V, 10 V, 30 V, 100 V und 300 V. Die Verstärkerwiedergabe schwankt nicht mehr als $\pm 3\%$ von 15 Hz bis 150 kHz und $\pm 5\%$ von 5 Hz bis 500 kHz.

HOCHLEISTUNGS-AUDIOFREQUENZ-OSZILLATOR

(Abbildungen auf Seite 237)

Dieser Oszillator ist ein *RC*-Abgleichungstyp für den Frequenzbereich von 20 Hz bis 20 kHz. Ein Merkmal des Oszillators ist die hohe Ausgangsleistung (6 W), die einer Lastimpedanz von 15Ω oder 600Ω angepasst ist, d.h. ausreichende Kraft zum direkten Betrieb von Lautsprechern oder kleinen Schwingungsgebern. Die Genauigkeit beträgt $\pm 1\%$ im Bereich von 200 Hz bis 2 kHz mit einer Genauigkeit von $\pm 2\%$ in anderen Bereichen.

EE 8758 *für weitere Einzelheiten*

ELECTROTHERMAL ENGINEERING LTD (135)

270 Neville Road, London E.7

PRÄZISIONS-HARTLÖTEGERÄT

(Abbildungen auf Seite 237)

Dieses Gerät wurde entwickelt, um ein leicht bedienbares, tragbares und kräftiges Aggregat zur Verwendung mit Silberhartlötelegierungen zu schaffen. Es ist besonders vorteilhaft für das Hartlöten von Widerstandsdrähten, wie z.B. Nickelchrom und verwandte Legierungen, wo Weichlot entweder schwierig zu verwenden ist oder einen zu tiefen Schmelzpunkt für die betreffende Arbeit aufweist, und wo Punktschweissen eine zu schwache Naht ergibt.

Obwohl es in der Hauptsache für Feinarbeit entworfen ist, von 0,025 mm Durchmesser, kann das Aggregat zum Verlöten von Draht mit einer Stärke bis 10 swg (3,25 mm) Durchmesser verwendet werden.

Der Kraftverbrauch des Aggregats richtet sich nach der Steuerungseinstellung; der Höchstverbrauch beträgt ungefähr 500 W. Das Lötmittel wird vermittels eines sich abnützenden Elements erwärmt, das sich leicht ersetzen lässt. Behälter für Schmelzmittel und für die Lösungen zur Schmelzmittelentfernung sind in bequemer Reichweite des Bedienungsmanns angebracht, wobei die Lösungen erwärmt werden. Das Gerät ist zum Betrieb an jeder beliebigen Wechselstromquelle von 200 bis 250 V ohne Einstellung geeignet, und an weiteren Weshselstromquellen auf Sonderbestellung.

EE 8759 *für weitere Einzelheiten*

FORTIPHONE LTD (152)

92 Middlesex Street, London E.1.

SUBMINIATUR-POTENTIOMETER

Das Subminiatur-Potentiometer und

einpolige Schalter Typ VS 32 besitzt ausserordentlich kleine Abmessungen und ist zur Verwendung in am Kopf befestigten Hörilfen usw. bestimmt. Es ist für Frontund Seitenbefestigung entworfen, wird vermittels einer Mittenschraube und Mutter befestigt und hat einen Gesamtdurchmesser von 8 mm, und eine Tiefe von 3,8 mm. Es kann in Widerstandswerten zwischen 1 kΩ und 3 MΩ und einem Kraftnennwert von 50 mW geliefert werden.

Das Vorwähl-Potentiometer Typ PSR steht für herkömmliche oder gedruckte Schaltungszwecke zur Verfügung. Der Gesamtdurchmesser beträgt 11,4 mm und die Gehäusetiefe beträgt 2 mm. Der Widerstandsbereich und Kraftnennwert sind dieselben wie für den Typ VS 32.

Der Typ PSS ist ein Vorwähl-Gleitdrahtpotentiometer einfacher Konstruktion und Anordnung mit einer Gesamtlänge von 28 mm und Gesamtbreite von 7,2 mm. Der Widerstandsbereich beträgt 1 kΩ bis 3 MΩ und der Kraftnennwert 50 mW.

EE 8760 *für weitere Einzelheiten*

GEO. L. SCOTT & CO. LTD (133)

Cromwell Road, Ellesmere Port, Cheshire

SCHICHT- UND RINGKERNE

Eine Reihe von Subminiatur-Transformatorschichtgernen steht nunmehr zur Verfügung mit einen Gestamtabmessungsbereich von 25 mm × 19 mm abwärts auf 9,5 mm × 9.5 mm, zum Einsatz in Transistorschaltungen, und überall da, wo die Platzfrage eine Rolle spielt.

Torus-Ringkerne, unter Verwendung von vermittels des Pulvermetallurgieverfahrens hergestellten Nickel-Eisen-Werkstoffen stehen auf keramischen Spulen mit hoher Wärmebeständigkeit, in RCL 193 Grössen auf welche Wicklungen direkt aufgebracht werden können, zur Verfügung. Die Beständigkeit der Kennwerte und die bei Verwendung dieser Legierungen erzielbaren hohen Leistungen machen sich bemerkbar, wo z.B. abgestimmte Paare für Magnetverstärker benötigt werden. Torus-Ringkerne aus Silizium-Eisen-Werkstoffen in Stärken von 0,33, 0,1 und 0,05 mm sind ebenfalls sowahl gemäss RCL 193 als auch nach Angaben der Kunden erhältlich, und zwar für NF- und HF-Kraftquellen.

Schnellschaltende und hochpermeable Bandkerne bis zu einer Mindeststärke von 0.008 mm werden ebenfalls auf keramischen Spulen für Schaltmatrizen und andere Hochleistungs-Magnetgeräte hergestellt.

EE 8761 *für weitere Einzelheiten*

HINCHLEY ENGINEERING CO. LTD (91)

Pans Lane, Devizes, Wiltshire

ISOLIERTRANSFORMATOREN

(Abbildungen auf Seite 238)

Neben einem grossen Bereich von Transformatoren für Radio- und Fernsehzwecke u.s.w. wird ausserdem eine Reihe von Standard-Isolier- und Einspulentransformatoren zum Einsatz in Laboratorien, Werkstätten und Wartungsabteilungen zur Schau kommen. Diese Transformatoren sind alle mit abgesicherten Schnellumschalt-Ausgangsspannungswählern ausgestattet, und sind wahlweise mit 'Mycalex' Schnappklemmmen lieferbar.

Angesichts der ansteigenden Zahl von elektrischen Haushaltsgeräten zum Einsatz in Werkstatt und Garten sind zwei Ausführungen des Hinchley 'Sicherheits'-Kastens, die tragbare in kräftigen Stahlkästen untergebrachte Isoliertransformatoren sind—(entweder 1 : 1 Verhältnis zur Isolierung von bestehenden Geräten; oder, vorzugsweise, 110 V oder 55 V Ausgang mit Zentralpunkt geerdet) von besonderem Interesse.

Dieselben wurden speziell für Grossmengenfertigung entworfen, so das die Endkosten für den Verbraucher die Sicherheit zu einer wirtschaftlicheren Notwendigkeit gestalten wird.

EE 8762 *für weitere Einzelheiten*

JAMES A. JOBLING & CO. LTD (158)

Wear Glass Works, Sunderland

METALLOXYD-WIDERSTÄNDE

Diese Hochpräzisions-, Hochstabilitäts-Widerstände sind aus 'Pyrex' Stäben hergestellt, auf welche ein dünner Belag von Metalloxyd aufgeschmolzen wird. Der Kern und der Belag sind untrennbar undurchlässig und stabil unter den allerungünstigsten Bedingungen und ihr Rauschpegel ist ausserordentlich niedrig.

Die Standardtoleranz dieser Widerstände beträgt 1% und sie werden vorzugsweise gemäss den genauen Anforderungen der Benützer an Stelle von festgesetzten Werten hergestellt, und zwar in einem Bereich von 100Ω und 250 kΩ, und mit Nennleistungen von 0,5 W bis 2 W.

EE 8763 *für weitere Einzelheiten*

LABGEAR LTD (110)

Willow Place, Cambridge

ZERSTÖRUNGSFREIES BLITZPRÜFGERAT

(Abbildungen auf Seite 238)

Dieses Gerät ist zum Prüfen der Isolierung von Schaltungen und Bauelementen wie Kondensatoren mit niedrigen Spannungen entworfen.

Das Standardmodell gibt Prüfspannungen von 250 V, 500 V, 1 kV, 1,5 kV und 2 kV. Ein Merkmal des Geräts ist, dass es das Prüfen eines Werkstücks mit einer über seiner normalen Betriebsspannung liegenden Spannung ermöglicht, so dass im Falle einer Störung das Prüfstück vollkommen unbeschädigt bleibt und bei einem niedrigeren Potential wieder verwendet werden kann, wodurch es möglich gemacht wird, Werkstücke ohne Zerstörungsgefahr zu sortieren.

EE 8764 *für weitere Einzelheiten*

LUSTRAPHONE LTD (111)

St. George's Works, Regents Park Road, London. N.W.1

STEREOFONISCHES MIKROFON

(Abbildungen auf Seite 238)

Dieses Mikrofonaggregat besteht aus zwei identischen Bändchen-Geschwindigkeitsmikrofonen, die senkrecht untereinander montiert sind. Die obere Hälfte der Einheit, die eines der Mikrofone enthält, ist so angeordnet, dass sie durchgehend durch einen Winkel von ungefähr 100° von der 'untereinander' Position gedreht werden kann, wodurch ein Höchstmass an Einstellungsmöglichkeiten für jeden Betriebsfall (stereofonische Übertragung, Tonaufnahmen, Senden, Tonverstärkung u.s.w.) erzielt werden kann. Die untere Hälfte des Mikrofons ist ortsfest angebracht.

Das Mikrofon ist mit einem Schalteraggregat verbunden, und zwar vermittels eines Vierwege-Schraubenverbinders. Der Vorteil dieser Methode besteht darin, dass der empfindliche Mikrofonkopf im Nu abgenommen und an einem sicheren Ort aufbewahrt werden kann, ohne Störungen in der Anordnung und den Anschlüssen zum Hauptgerät zu verursachen.

Ein besonders konstruierter Schalter mit drei Positionen wird geliefert:

(1) 'Aus' Position in der Mitte.

(2) Stereo-Anschluss am äussersten Ende einer Seite.

(3) Die andere Position am äussersten Ende der anderen Seite verbindet die beiden Bändchen-Aggregate in Reihenanschluss und verwandelt dadurch das Mikrofon in ein Geradeaus-Bändchengeschwindigkeitsmikrofon mit hoher Empfindlichkeit und einer 'Achter'-Richtcharakteristik.

An der Rückseite des Schalteraggregats ist ein Phasenumkehrschalter vorgesehen, damit zur Erzielung der richtigen Phasensteilung an jedem Gerät augenblickliche Umschaltung gegeben ist.

Wenn das Mikrofon, wie unter (3) oben beschrieben, in seine Stellung verbracht wird, und beide Einheiten sich 'untereinander' befinden, so bewirkt der Phasenumkehrschalter die Umwandlung des Mikrofons in ein rauschunterdrückendes Mikrofon von beträchtlicher Leistungsfähigkeit, das lediglich auf eine sich in nächster Nähe befindliche Schallquelle reagiert.

In Verbindung mit der wechselseitigen Position des Phasenschalters ergibt diesbezügliches Anordnen der Mikrofonaggregate in Zwischenstellungen eine Vielfalt von Richtwirkungen, die sich unter schwierigen akustischen Bedingungen als vorteilhaft erweisen können.

Die Impedanz jedes Mikrofonaggregats im Standardmodell beträgt 20Ω und die Gesamtimpedanz des auf Reihenposition geschalteten kombinierten Geräts beträgt 40Ω.

EE 8765 *für weitere Einzelheiten*

McMURDO INSTRUMENT CO. LTD (32)

Victoria Works, Ashtead, Surrey

RÖHRENFASSUNGEN- UND VERBINDER

Eine neue Reihe von schreibfederartigen Röhrenfassungen unter Verwendung von unbearbeiteten Einzelgiessformen für sowohl die B9A und B7G Version für gedruckte und hergebrachte Schaltungen steht zur Verfügung. Dieselben sind zum Aufnehmen von Röhren-Federhalterungen entworfen, deren Befestigung zusammenhängend mit dem Metallsockel der Röhrenfassung ist.

Darüber hinaus können die Röhren durch billige aufgerollte Abschirmabdeckungen geschützt werden, die eng um den Röhrenkolben anliegen und durch die Federhalterung festgehalten und elektrisch geerdet werden.

Die 'Pakonector' Reihe von Steckern und Sockeln wurde erweitert durch die Einführung einer 8-Wegeversion, die eine der Röhrenfassung B7G ähnliche Grösse aufweist. Derselbe Kabelanschluss und dieselbe Verkapselung wie bei der ursprünglichen 18-Wegeausführung stehen für die 8-Wegeversion zur Verfügung.

EE 8766 *für weitere Einzelheiten*

MULLARD LTD (77, 67)

Mullard House, Torrington Place, London WC1

110° Kathodenstrahlröhren

Grössen von sowohl 534 mm als auch 432 mm werden hergestellt, und zwar unter den Typennummern AW53-88 und AW43-88.

Die neuen 110° Röhren sind beträchtlich kürzer als die zur Zeit zur Verwendung kommenden entsprechenden Röhrengrössen, wodurch sich Einsparungen in den Abmessungen zwischen Vorder- und Rückseite bei Fernsehempfängertruhen ermöglichen lassen.

Beim 534 mm Schirm ist die 110° Röhre über 216 mm kürzer als ihr 70° Gegenstück, und 127 mm kürzer als eine 90° Type. Die 432 mm 110° Röhre weist Verminderungen von ungefähr 165 mm bezw. 76 mm gegenüber den entsprechenden Grössen der 70° und 90° Typen auf.

Die neuen Röhren besitzen Hochleistungs-Phosphore mit Metallhinterlegung, und ergeben ausgezeichnete Helligkeit und Kontraste bei starken Innenbeleuchtungen. Sie sind elektrostatisch fokussiert, und umfassen die 'gerade' Elektronenkanonen-Type, die einen Ionenfallen-Magnet überflüssig gestaltet.

EE 8767 *für weitere Einzelheiten*

PAINTON & CO. LTD (29)

Kingsthorpe, Northampton

Miniatur-Hochstabilitätswiderstände

Diese ausserordentlich kleinen Hochstabilitäts-Kohlenstoffwiderstände (Typ 70), messen lediglich ungefähr 9,5 × 2,5 mm, und sind deshalb zur Verwendung in Miniatur-transistorisierten Geräten und insbesondere zum Einsatz bei gedruckten Schaltungsplatinen geeignet. Die Leitungen besitzen einen Vorüberzug von Zinn, um Tauchlöten zu erleichtern. Der Widerstandsbereich umfasst Standardwerte von 5Ω bis 250 kΩ bei einer Toleranz von 5%, aber wenn erforderlich stehen Toleranzen von 1% und 2% für den grösseren Teil des Bereichs zur Verfügung.

EE 8768 *für weitere Einzelheiten*

Gedruckte Schaltungsverbindungen

(Abbildungen auf Seite 239)

Eine Anzahl von Verbindungsgliedern für gedruckte Schaltungen wurden eingeführt, einschliesslich eines kräftig konstruierten 15-gängigen Steckers und Steckerleiste mit Stahlstiftbestückung. Zusammen mit den meisten 'Painton' Verbindungsgliedern für gedruckte Schaltungen beruhen die Kontakte auf einem 2,54 mm Modul.

Ebenfalls in der Fertigung befindet sich eine Verbindungsleiste mit bis zu 60 Kontakten in drei Gruppen von 20 mit 2,54 mm Mitten; die Kontakte sind von einzigartiger und patentierter Konstruktion, und Prüfungen haben erwiesen, dass diese Kontaktform eine lange Standzeit der Platineneinsteckung gestattet.

EE 8769 *für weitere Einzelheiten*

Rechtwinklige Röhrenfassungen

(Abbildungen auf Seite 239)

Diese Bauelemente wurden dafür entworfen, waagrechten Einbau von Röhren zu gestatten, und auf diese Weise die Fertigung von gedrungeneren gedruckten Schaltungsmontagen zu erleichtern und Wärmeableitanordnungen zu vereinfachen.

Die Kontakte bestehen aus silberüberzogenem Berylliumkupfer und sind lang genug, um sich für die meisten gewöhnlich zur Verwendung kommenden Stärken von gedruckten Schaltungsplatinen zu eignen. Die korrekten Einführungs- und Abnahmedrücke werden bei einem Mindestmass an Kontaktwiderstand aufrechterhalten.

EE 8770 *für weitere Einzelheiten*

STEATITE AND PORCELAIN PRODUCTS LTD (76)

Stourport-on-Severn, Worcestershire

Klemmenplomben

(Abbildungen auf Seite 239)

Im Anschluss an die letztjährige Vorankündigung wird die neue Reihe von Klemmenplomben die entsprechend den Anforderungen der 'Streitkräfte'—Norm RCL 331 entworfen wurden, zur Schau kommen. Dieselben sind aus hochwertigen Alumina-Keramikwerkstoffen hergestellt, und die Ableitungen sind fest in ihrer Stellung vermittels Schmelzglasur eingeschmolzen. Bänder mit Metallüberzug, die vermittels eines für den erforderlichen Gütegrad genehmigten Nickel-überzugverfahrens hergestellt werden, gestatten Weichlöten direkt in Metallhülsen für alle hermetischen Abdichtungsverwendungen.

Die Standardreihen von Plomben mit Nickelüberzug, die für die H1 Gütegruppe genehmigt sind, stehen noch ab Lager zur Verfügung, und eine Musterabteilung stellt Grundtypen von neuen Entwürfen in kürzester Zeit her. Besonderes Augenmerk wurde auf verbesserte Vorrichtungen für die Montage von metallüberzogenen keramischen Erzeugnissen mit geeigneten Metallbauteilen gerichtet, um eine grosse Vielfalt von Verwendungszwecken, angefangen von Miniaturableitungen bis zu grossen Hochspannungs-Kondensatorklemmen, zu erfassen.

EE 8771 *für weitere Einzelheiten*

SWIFT LEVICK & SONS LTD (156)

Clarence Steel Works, Sheffield, 4

Grosse Dauermagnete

(Abbildungen auf Seite 240)

Sehr grosse und stabile Dauermagnete werden jetzt mit Feldern hoher Homogenität für magnetische Kernresonanzzwecke hergestellt. Die Fotografie zeigt einen solchen Magnet, der für Forschungszwecke für der Universität von Leeds etworfen und hergestellt wurde. Dieser Magnet wiegt 2,350 kg, und seine Gesamtabmessungen betragen 1 210 × 1 070 × 610 mm; die Dauermagnetisierung geschieht vermittels von Clumax-Legierung, der stabilste und stärkste Magnetwerkstoff, der zur Zeit handelsmässig auf der ganzen Welt hergestellt wird. Der Magnet hat eine Spaltfeldstärke von 9 700 Oe innerhalb eines Spaltraums von 36 mm lang × 254 mm Durchmesser, und die Feldgleichheit ist besser als ein Teil in 10^6 innerhalb eines mittleren Raums von 1 cm^3. Besondere mechanische Justierungen sind an einem der eisernen Polteile vorgesehen, wodurch eine beschränkte Bewegungsfreiheit der Poloberfläche in jeder beliebigen Ebene für das endgültige Einstellen der Spaltfeldstärke gestattet ist.

Ein Modell dieses Magnets im Masstab 1:6 kommt zur Ausstellung.

EE 8772 *für weitere Einzelheiten*

TELEPHONE MANUFACTURING CO. LTD (21)

Hollingsworth Works, London, S.E.21

Mehrfachrelais

(Abbildungen auf Seite 240)

Das neue zehngängige Mehrfachrelais wird zur Schau kommen. In diesem neuen Relais kommen viele einzigartige und neuartige Methoden zur Verwendung, die sowohl niedere Kosten als auch erhöhte Zuverlässung mit sich bringen. Das Mehrfachrelais TMC verwendet einen gemeinsamen Steg mit Montagestützen, und ist zum Ersatz von Relais der Telefontype in Steuerschaltungen bestimmt, wo der Kostenpunkt an erster Stelle steht.

Kontaktfedern aus Phosphorbronze, mit Silber-Bimetallkontakten unter Anwendung der Kreuzradienmethode, und automatische Maschineneinstellung und Kartenbetrieb gewährleisten einen stetigen Kontaktdruck, wie er bisher nur in teuren Präzisionsrelais zur Verfügung stand. Eine leistungsfähige magnetische Schaltung gewährleistet, dass magnetische Einwirkungen zwischen Relais so gut wie nicht auftreten, während die Montagestützen mit gemeinsamem Steg die Anzahl der Einzelbestandteile beträchtlish verringern und die Notwendigkeit für getrennte Montageteile überflüssig gestalten.

Der Preis dieses Mehrfachrelais ist so gehalten, dass es sich oft wirtschaftlicher erweist, eine zehngängige Einheit an Stelle einer geringeren Anzahl von herkömmlichen Relais anzubringen.

EE 8773 *für weitere Einzelheiten*

WHITELEY ELECTRICAL RADIO CO. LTD (81)

Radio Works, Victoria Street, Mansfield, Nottinghamshire

KABELPRÜFGERÄT

(*Abbildungen auf Seite 240*)

Dieses Prüfgerät bestimmt die Lage von Bruchstellen oder Kurzschlüssen in ummantelten vieladrigen Leitungskabeln bis auf den Bruchteil eines Zolls. Reparaturen können dann mit einem Mindestmass von Beschädigung des Kabelmantels ausgeführt werden.

Das Betriebsprinzip ist wie folgt: Im Falle eines Leitungsbruchs wird eine Tonfrequenzspannung zwischen einem Ende der Leitung und Erde angelegt; das elektrostatische Feld zwischen dem unter Strom stehenden Abschnitt der Leitung und Erde lässt sich dann durch das Entlangführen einer kapazitiven Sonde bestimmen. Die Sonde ist an den Eingang eines trgbaren Transistorverstärkers angeschlossen, der ein Kopfhörergerät speist und auf diese Weise eine hörbare Anzeige der Bruchstelle ergibt. Alle anderen Leitungen im Kabel müssen geerdet sein, mit Ausnahme der sich unter Prüfung befindlichen.

Eine induktive Sonde wird an Stelle der kapazitiven Sonde zum Auffinden von Kurzschlüssen zwischen Leitungen verwendet. Der Oszillator speist ein Signal in die Leitung nur bis zu der Stelle, wo der Kurzschluss auftritt, und die Sonde bestimmt dann das elektrostatische Feld bis zu dieser Stelle.

Das Prüfschaltbrett ist mit Drehschaltern bestückt, um das Wählen von Einzelleitungen im Kabel zu ermöglichen. Alle Leitungen mit Ausnahme der unter Prüfung stehenden sind geerdet. Steckdosen zur Aufnahme der Kabelstecker sind angebracht. Zur anfänglichen Bestimmung einer schadhaften Leitung ist eine leuchtenähnliche Stetigkeitsprüfvorrichtung vorgesehen.

Der Röhrenoszillator ist batteriebetrieben und besitzt eine Ausgangsleistung von 10 mW mit einer Ausgangsimpedanz von 300Ω. Seine Frequenz beträgt 1 kHz ±2½%. Je nach der Einstellung des Potentiometers kann die Art der Schwingung entweder andauernd oder unterbrochen sein.

Der Transistorverstärker wird mittels einer im Gehäuse untergebrachten 4½ V-Batterie betrieben und hat eine Gesamtverstärkung von 70 dB, bei einer nichtinduktiven Ausgangslast von 300Ω.

Die induktive Sonde ist auf 1 × kHz ±2½% abgestimmt.

EE 8774 *für weitere Einzelheiten*

THE ZENITH ELECTRIC CO LTD. (8)

Zenith Works, Villiers Road, Willesden Green, London, N.W.2

REGELTRANSFORMATÖREN

(*Abbildungen auf Seite 240*)

Ein neues 'Variac' Modell 50-B/2B mit Doppelausgang wird ausgestellt werden. Dasselbe ist für einen Kraftanschluss von 230 V 50 Perioden entworfen, wodurch sich eine zwischen null und 270 V regelbare Leistung ergibt, mit einem Höchstnennstrom von 20 A, was den Gesamtwert darstellt, der den beiden Bürsten entnommen werden kann, aber die Last lässt sich natürlich verteilen, um Anwendungsforderungen wie z.B. 1-12 A für Ausgang Nr.1, und 8 A für Ausgang Nr.2 zu entsprechen. Die Doppelbürsten, Doppelausgang 'Variacs' stehen seit einiger Zeit in den Reihen '100' und 'V30H', und ausserdem in dem kleineren Torusbereich in der Reihe '200' zur Verfügung.

Diese kleineren Typen wurden nunmehr unter Verwendung von Bauelementen der 'V5' Reihe neu entworfen, wodurch die Fertigung eines kompakteren Typs ermöglicht wird.

EE 8775 *für weitere Einzelheiten*

ERDUNGSSCHLEIFEN-PRÜFGERÄT

Dasselbe ist in einen Metallschrank, komplett mit Traggriff und einer Tasche für ein Zuleitungskabel und Zwischenstecker, untergebracht. Es ist für das Vornehmen aller in der I.E.E. Vorschrift Nr.507 festgelegten Prüfungen, und zum Betrieb von einer Standard-Einphasenquelle entworfen, mit einer Ausgangsleistung von 25 A und jeder beliebigen Spannung zwischen null und 40 V.

Ein 'Zenith' Abwärtstransformator mit Doppelwicklung kommt mit einen 'Variac' über der Primärwicklung zur Verwendung, und ausserdem ein Doppelbereich-Gleichrichtervoltmeter zusammen mit einem Ammeter; wir liefern je eine 13 A und 15 A Dreikontaktstift-Steckdose an der Vorderseite des Gehäuses für Erdkernprüfungen tragbarer Geräte. Das 40 V Gerät ist für die meisten Prüfarbeiten erhältlich, aber 6 V is mit eingeschlossen, das sich dies für kurze Erdungsschleifen wie z.B. bei tragbaren Werkzeugen und anderen ähnlichen Geräten als geeigneter erweisen mag.

Ein Verbindungsglied zum Prüfen von ortsfesten Geräten ist eingebaut, für welches das 6 V verwendet werden soll, da die Erdleitung verhältnismässig kurz ist und 0,1Ω, was einen Wert von ungefähr 3 V darstellt, nicht übersteigen darf.

EE 8776 *für weitere Einzelheiten*

Berichtisung

Wir bedauern, dass in der Beschreibung des 'Magnetband-Speichers' der Firma Epsylon Industries Ltd auf Seite 188 unserer März-Ausgabe die Bandgeschwindigkeiten unrichtig angegeben wurden. Die richtigen Werte sind:—

Standard-Bandgeschwindigkeit — 254 cm/Sek.

Zukünftige Bandgeschwindigkeit — 508 cm/Sek.

Schnellaufspulung—1 016 cm/Sek.

Zusammenfassungen der Hauptartikel

Lineare Mehranzapfstellen-Potentiometer mit belasteten Ausgängen von K. C. Garner

Zusammenfassung des Artikels auf Seite 192-199

Eine Analyse von Mehranzapfstellen-Potentiometern wird hier gegeben, und zwar für die Type, wo Shunt-Widerstände zwischen angrenzenden Anzapfstellen angeschlossen sind, und beim Vorhandensein eines Ausgangsbelastung-Widerstands.

Bei Verwendung eines Allgemeinnetzes mit entsprechenden ziffernmässigen Einsatzwerten, die aus der in Frage stehenden Potentiometeranordnung ermittelt werden, gestaltet sich die Auswertung einfach und systematisch.

Ein Verfahren für den Entwurf von Potentiometern zum Erzeugen von nicht-linearen Funktionen wird vorgeschlagen.

Aufschlaggeräusch-Analyse von F. M. Savage

Zusammenfassung des Artikels auf Seite 200-203

Geräte für das Messen und die Analyse von anhaltenden Geräuschen haben es ermöglicht, Normen zum Vermeiden von Gehörbeeinträchtigung bei Personal aufzustellen, das ständigen Geräuschen ausgesetzt ist. Die Entwicklung von Analysiergeräten für Aufschlaggeräusche dürfte die erforderlichen Unterlagen zur Behandlung der sogar noch gefährlicheren Aufschlaggeräusche erbringen. In diesem

Artikel wird ein kleines tragbares Gerät zum Messen von kury anhaltenden Geräuschen beschrieben. Das Gerät kann auch für andere Zwecke verwendet werden, wie z.B. das Messen von Spitzenbeschleunigungen, wenn es mit einem geeigneten Beschleunigungsmesser verwendet wird.

Transistorverstärker für Magnetband-und Trommelabspielgerät von A. E. Bachmann

Zusammenfassung des Artikels auf Seite 213-217

Magnetbänder- und Trommeln werden häufig als Speicheröder Übertragungselemente verwendet, wobei die Information in digitaler Form enthalten ist. Diese Information lässt sich durch Abspielen des Bands oder der Trommel durch einen Magnetkopf und Verstärker wieder zur Verfügung stellen. Der Artikel bespricht die mit der Konstruktion eines solchen Verstärkers verbundenen Probleme, wenn lediglich Transistoren verwendet werden. Ein Grundtyp wurde geprüft, der von einem Abspielkopf betätigt wird und eine Arbeitsstrom-Ausgangsspannung von mehr als 2 mV aufweist. Er ist in der Lage, drei bistabile Multivibratoren in Parallelschaltung innerhalb eines Frequenzbereichs von 2 bis 400 kHz und einem Temperaturbereich von 0°C bis +70°C zu betätigen. Die Anstiegzeit des Ausgangsimpulses beträgt weniger als 1 μSek.

Ein einfacher NF-Oszillograph von J. F. Young

Zusammenfassung des Artikels auf Seite 218-220

Die gewöhnlichen Methoden der Stabilisierung von RC-Oszillographen sind frequenzempfindlich. In dem beschriebenen einfachen Oszillograph begrenzt eine Zener-Diode die Amplitude and die sich ergebenden Oberwellen werden vermittels eines Wählkreises, der ausserdem die Schwingungsfrequenz steuert, ausgefiltert. Eine annehmbare Wellenform lässt sich deshalb selbst bei sehr niederen Frequenzen erzielen.

Eine stroboskopische Methode zum Vornehmen von Frequenzwiedergabe-Messungen an kleinen elektro-mechanischen Vorrichtungen von M. Sheperdson und R. Walters

Zusammenfassung des Artikels auf Seite 220-221

Eine Methode zum Vornehmen von Frequenzwiedergabe-Messungen an kleinen elektromechanischen Vorrichtungen unter Verwendung eines beweglichen Mikroskops und einer stroboskischen Lampe wird beschrieben. Abgesehen von einem einfachen Impulsverstärker- und Phasenverschiebungsnetz, das beschrieben wird, steht die erforderliche Apparatur in den meisten gut ausgestatteten elektrischen Laboratorien zur Verfügung. Die Methode erfordert kein Einspannen des Prüflings und gestattet das Vornehmen von Amplitudenmessungen weit unter 0,025 mm.

Regulierte Stromquellen von D. J. Collins und J. E. Smith

Zusammenfassung des Artikels auf Seite 222-226

Regulierte, oder stabilisierte Gleichstromquellen werden für eine Reihe von Zwecken benötigt. In diesem Artikel werden verschiedene Typen von Regulatoren kurz besprochen und der Reihenstabilisator mit geschlossener Schleife wird etwas ausführlicher behandelt. Die Konstruktionsprobleme dieser Aggregattype werden untersucht und der Schleifenstabilität wird besondere Beachtung geschenkt. In einem Anhang wird der Entwurf einer typischen Einheit in allen Einzelheiten gezeigt.

Telemetrische Verfahren unter Verwendung des Grundprinzips der gasgefüllten Stufenröhre von Yoshisuke Hatta

Zusammenfassung des Artikels auf Seite 227-229

Ein telemetrisches Verfahren zum Ermitteln von mechanischer Verschiebung unter Verwendung des Grundprinzips der gasgefüllten Stufenröhre wird beschrieben. Da es sich um ein elektronisches Gerät handelt, besitzt es eine vorteilhafte hohe Betriebsgeschwindigkeit, und ist deshalb in mancher Beziehung elektromagnetischen Systemen, denen beträchtliche Trägheit zu eigen ist, überlegen.

Feste elektrolytische Tantalkondensatoren von R. Aries

Zusammenfassung des Artikels auf Seite 230-231

Feste elektrolytische Tantalkondensatoren haben bereitwillige Aufnahme für die Verwendung in miniaturisierten Geräten gefunden. In diesem Artikel wird eine kurze Beschreibung des Fertigungsverfahrens zusammen mit den Ergebnissen von Lebensdauerprüfungen und Temperaturkennwerten gegeben.